図面の簡単な説明

図１は、ＯＦＤＭ伝送システムにおいて使用される既存の送信装置の構成図である。

図２は、既存のＯＦＤＭ伝送システムにおける伝送信号の例である。

図３は、ＯＦＤＭ伝送システムにおいて使用される既存の受信装置の構成図である。

図４は、マルチパスを説明する図である。

図５は、複数の移動機を収容する基地局を示す図である。

図６は、本発明の実施形態の送信装置の構成図である。

図７は、本発明の実施形態の受信装置の構成図である。

図８および図９は、実施形態のＯＦＤＭ伝送システムにおける伝送信号の例である。

図１０は、ガード区間について説明するための図である。

図１１は、副搬送波変調器により実行される逆フーリエ変換を説明する図である。

図１２は、ガード区間を挿入する処理を説明する図である。

図１３は、ガード区間を挿入する処理を実現する構成の実施例である。

図１４は、受信波からガード区間を削除する処理を実現する構成の実施例である。

図１５は、第１の実施例の送信装置の構成図である。

図１６は、第１の実施例の受信装置の構成図である。

図１７は、第１の実施例の通信システムにおける伝送信号を模式的に示す図である。

図１８は、第２の実施例の送信装置の構成図である。

図１９は、第２の実施例の受信装置の構成図である。

図２０は、第２の実施例の通信システムにおける伝送信号を模式的に示す図である。

図２１は、第３の実施例の送信装置の構成図である。

図２２は、第３の実施例の受信装置の構成図である。

図２３は、図２２に示す遅延差検出部の一例の構成図である。

図２４は、遅延差検出部の動作を説明する図である。

図２５は、最大伝送遅延差を検出する実施例である。

図２６は、第４の実施例の送信装置の構成図である。

図２７は、第４の実施例の受信装置の構成図である。

図２８は、図２７に示す距離推定部の一例の構成図である。

図２９は、第５の実施例の送信装置の構成図である。

図３０は、第５の実施例の受信装置の構成図である。

図３１は、図３０に示すタイミング生成部の一例の構成図である。

図３２は、第６の実施例の送信装置の構成図である。

図３３は、第６の実施例の受信装置の構成図である。

図３４は、図３３に示すタイミング生成部の一例の構成図である。

図３５は、第７の実施例の送信装置の構成図である。

図３６は、第７の実施例の受信装置の構成図である。

図３７は、図３６に示す遅延差検出部の動作を示すフローチャートである。

図３８は、第８の実施例の送信装置の構成図である。

図３９は、第８の実施例の受信装置の構成図である。

図４０は、図３９に示す距離推定部の動作を示すフローチャートである。

発明を実施するための最良の形態

本発明の実施形態について図面を参照しながら説明する。以下では、セルラ

通信システムにおいて直交周波数分割多重・符号拡散（ＯＦＤＭ－ＣＤＭ）伝送方式が利用されるものとする。具体的には、例えば、図５に示す基地局と移動機との間の信号伝送のためにＯＦＤＭ－ＣＤＭが利用されるものとする。

図６は、本発明の実施形態の送信装置の構成図である。なお、この送信装置は、図５においては、例えば、基地局装置に相当する。また、この送信装置は、信号系列Ｓｉ および信号系列Ｓｊ を多重化して出力するものとする。ここで、信号系列Ｓｉ および信号系列Ｓｊ は、例えば、互いに異なる移動機へ送信すべき信号であってもよい。あるいは、信号系列Ｓｉ または信号系列Ｓｊ の中にそれぞれ複数の移動機へ送信すべきデータが時間多重されていてもよい。

この送信装置は、送信すべき信号系列毎に、拡散変調器（ＳＭＯＤ：Spread Modulator）１、副搬送波変調器（ＦＭＯＤ：Frequency Modulator）２、加算器（ＳＵＭ）３、ガード区間挿入器（ＧＩＮＳ：Guard Interval Insert Unit）２１、利得調整器（Ｇ）２２を備える。ここで、拡散変調器１、副搬送波変調器２、加算器３については、図１を参照しながら説明したものを使用することができる。すなわち、拡散変調器１は、ｍ個の入力端子を備えており、それらの入力端子には、シンボル周期Ｔとごに、同一のシンボル情報が並列に入力される。そして、拡散変調器１は、入力されたシンボル情報を信号系列Ｓｉ に対して予め割り当てられている拡散符号Ｃｉ を用いて変調し、その結果として得られるｍビットの拡散信号を出力する。なお、拡散符号Ｃｉ は、「Ｃi(1)」～「Ｃi(m)」から構成されており、直交符号列の中の１つの要素であるものとする。

副搬送波変調器２は、互いに異なる角周波数ω１～ωｍ を持ったｍ個の副搬送波を生成する。ここで、ω１, ω２, ω３, . . . ωｍ の角周波数間隔Δωは、シンボル周期Ｔの逆数により定義される一定の値であり、下記の式により表される。

$\Delta\omega = 2\pi\Delta f = 2\pi / T$

また、副搬送波変調器２は、拡散変調器１から出力される拡散信号を用いて

m個の副搬送波を変調する。具体的には、例えば、角周波数ω1を持った副搬送波は、「Ci(1)」が乗算されたシンボル情報により変調され、角周波数ωm を持った副搬送波は、「Ci(m)」が乗算されたシンボル情報により変調される。なお、副搬送波変調器2の処理は、例えば、逆フーリエ変換演算により実現される。そして、副搬送波変調器2から出力される各副搬送波は、加算器3により合成される。

ガード区間挿入器２１は、シンボル毎に、加算器3から出力される合成信号に対して、ガード区間（Guard Interval）を挿入する。ここで、このガード区間は、無線伝送路のマルチパスによる影響を排除するために挿入される。なお、図1に示した既存の送信装置のガード区間挿入器4は、予め固定的に決められたガード区間を挿入するが、実施形態のガード区間挿入器２１は、送信装置と受信装置との間の通信状態に応じて決められるガード区間を挿入する。なお、ガード区間の長さは、ガード区間制御部（ＧＩＮＳＣＮＴ：Guard Interval Control Unit）２３により、信号系列ごとに決定される。

利得調整器２２は、例えば乗算器であり、ガード区間が挿入された信号に利得係数αを乗算する。これにより、送信すべき信号の振幅または電力が調整される。なお、利得係数αは、基本的に、信号系列ごとに挿入されるガード区間の長さに対応して決定される。

上述のようにして得られる各信号系列ごとの合成信号は、図1に示した既存の送信装置と同様に、加算器（ＡＤＤ）５により加算される。そして、加算器５の出力は、送信機（ＴＸ）６により所定の高周波信号に変換された後、アンテナ7を介して送信される。

このように、実施形態の送信装置では、送信すべき信号系列（Si、Sj）ごとに、送信装置と受信装置との間の通信状態に応じて決められるガード区間が挿入される。また、送信すべき信号系列（Si、Sj）ごとに、挿入されたガード区

間の長さに対応して送信信号の振幅または電力が調整される。

図７は、本発明の実施形態の受信装置の構成図である。ここでは、この受信装置は、図６に示す送信装置により送信された無線信号から信号系列Ｓｉを受信するものとする。なお、この受信装置は、図５においては、例えば、移動機に相当する。また、図７では、信号を受信するために必要な周波数同期機能、およびタイミング同期機能などは省略されている。

アンテナ１１により受信された信号は、受信機（ＲＸ）１２によりベースバンド信号Ｓrx に変換された後、副搬送波復調器（ＦＤＥＭ：Frequency Demodulator）１３によりm個の受信信号列に変換される。ここで、副搬送波復調器１３は、m個の入力端子を備えており、それらの入力端子には、同一のベースバンド信号Ｓrx が並列に入力される。そして、副搬送波復調器１３は、ベースバンド信号Ｓrx に対してそれぞれ角周波数ω1～ωm を持った周期波を乗算することにより、各副搬送波ごとに信号を復調する。なお、副搬送波復調器１３の処理は、例えば、フーリエ変換演算により実現される。

ガード区間削除器３１は、ガード区間制御部（ＧＣＮＴｉ：Guard Interval Control Unit）３２からの指示に従って、各受信信号列からそれぞれガード区間を削除する。なお、ガード区間制御部３２は、送信装置において信号系列Ｓｉに対して挿入されたガード区間の長さを認識しており、その値をガード区間削除器３１に通知する。したがって、ガード区間削除器３１は、送信装置で挿入されたガード区間を適切に除去することができる。

拡散復調器１５は、各受信信号系列を逆拡散するために、送信装置において使用された拡散符号と同じ拡散符号Ｃｉを各受信信号列にそれぞれ乗算する。そして、加算器１６を用いて拡散復調器１５から出力される各信号を加算することにより、信号系列Ｓｉが再生される。

図８および図９は、実施形態のＯＦＤＭ伝送システムにおける伝送信号の例

である。ここで、図８は、最大伝送遅延差の小さい位置にいる移動機（受信装置）へ送信すべき伝送信号を模式的に示しており、図９は、最大伝送遅延差の大きい位置にいる移動機へ送信すべき伝送信号を模式的に示している。なお、図８に示す伝送信号のシンボル周期が「Ｔ1」であるのに対し、図９に示す伝送信号のシンボル周期が「Ｔ2」であるが、これらの周期は互いに同じであってもよいし、互いに異なっていてもよい。

最大伝送遅延差の小さい位置にいる移動機へ信号を送信する場合は、図８に示すように、各副搬送波ごとに、シンボル周期Ｔ1 に対してガード区間Ｔg1 が挿入される。したがって、信号は、区間Ｔs1 を利用して伝送される。一方、最大伝送遅延差の大きい位置にいる移動機へ信号を送信する場合は、図９に示すように、各副搬送波ごとに、シンボル周期Ｔ2 に対してガード区間Ｔg2 が挿入される。したがって、信号は、区間Ｔs2 を利用して伝送される。そして、このとき、ガード区間Ｔg1 は、ガード区間Ｔg2 よりも短く設定される。すなわち、ガード区間の長さは、送信装置から受信装置へ信号が伝送されたときの最大伝送遅延差が大きくなると、それに応じて長くなる。

また、最大伝送遅延差の小さい位置にいる移動機へ信号を送信する場合は、図８に示すように、信号の送信電力は「Ｐ1」に制御される。一方、最大伝送遅延差の大きい位置にいる移動機へ信号を送信する場合は、図９に示すように、信号の送信電力は「Ｐ2」に制御される。ここで、電力Ｐ2 は、電力Ｐ1 よりも大きい。すなわち、信号の送信電力は、送信装置から受信装置へ信号が伝送されたときの最大伝送遅延差が大きくなると、それに応じて大きくなる。

続いて、ガード区間の挿入／除去について説明する前に、ガード区間そのものについて簡単に説明をする。

図１０は、ガード区間について説明するための図であり、受信装置が受信した信号の波形が模式的に示されている。ここで、実線ａは、受信装置に最初に

到着した信号（基準波）の波形を表し、破線ｂは、受信装置に到着した遅延信号（遅延波）の波形を表している。なお、図１０では、１つの遅延波のみが描かれているが、実際には、通常、２以上の遅延波が存在する。

図１０において、時刻Ｔ１以前は、基準波および遅延波がそれぞれ連続したサイン波なので、受信装置は、それらの合成波から対応するシンボル情報を再生することができる。しかし、シンボル情報が「＋１」から「－１」に変化したとき、あるいは「－１」から「＋１」に変化したときは、そのシンボル情報を伝送する信号の位相が転移する。図１０に示す例では、時刻Ｔ１において基準波の位相が転移しており、時刻Ｔ2において遅延波の位相が転移している。すなわち、この場合、時刻Ｔ１と時刻Ｔ2との間の期間では、基準波は位相転移後の情報を伝送しており、遅延波は位相転移前の情報を伝送していることになる。したがって、この期間は、一方の信号波が他方の信号波に対する干渉波となり、受信波からシンボル情報を適切に再生することができないことがある。

上記干渉による影響は、例えば、図１０に示す例では、受信波から信号を再生する際に、時刻Ｔ１と時刻Ｔ2との間の受信波を使用しないことにより回避される。そして、ＯＦＤＭ通信システムでは、この期間を含む所定の期間をガード区間として定義し、受信装置において信号再生が行われないようにしている。したがって、ガード区間の長さは、最初に到着する信号波と最後に到着する遅延波との遅延差（最大伝送遅延差）よりも大きく設定される必要がある。

ところが、上述したように、最大伝送遅延差は、送信装置と受信装置との間の距離などにより変化する。したがって、実施形態の通信システムでは、ガード区間の長さが最大伝送遅延差に対応して決定されるようになっている。

次に、送信装置においてガード区間を挿入する方法を説明する。ここでは、図６に示す副搬送波変調器２の処理は、逆フーリエ変換演算により実現されるものとする。

図１１は、副搬送波変調器２により実行される逆フーリエ変換を説明する図である。ここでは、シンボル周期を「Ｔ」、シンボル周期ごとに挿入されるガード区間を「Ｔg」、シンボル周期ごとの信号時間を「Ｔs（＝Ｔ－Ｔg)」とする。

副搬送波変調器２には、上述したように、拡散変調器１から出力されるｍ個の情報が入力される。ここで、各情報は、それぞれ対応する周波数の副搬送波に割り当てらる。すなわち、副搬送波変調器２は、周波数軸上に配置されたｍ個の信号を受ける。そして、この周波数軸上のｍ個の信号は、図１１に示すように、シンボル周期Ｔごとに実行される逆フーリエ変換により、時間軸上のｍ個のサンプルから構成される信号系列に変換される。このとき、時間軸上のｍ個のサンプルは、信号時間Ｔs内に配置される。

図１２は、ガード区間を挿入する処理を説明する図である。ガード区間挿入器２１は、信号時間Ｔs内に配置されたｍ個のサンプルを受け取ると、ガード区間Ｔgに相当する個数のサンプル成分を信号時間Ｔsの末尾から抽出し、それらを信号時間Ｔsの直前に複写する。図１２に示す例では、ガード区間Ｔgが３サンプリング時間に相当し、ｍ個のサンプル「１」～「ｍ」のうちから、「ｍ－２」「ｍ－１」「ｍ」が抽出されて信号時間Ｔs の直前に複写されている。そして、この複写により、シンボル時間Ｔ（＝Ｔg＋Ｔs）の時間軸上の信号系列が作成される。

図１３は、ガード区間を挿入する処理を実現する構成の実施例である。上述したように、副搬送波変調器２は、逆フーリエ変換器によって実現され、シンボル周期ごとに、周波数軸上のｍ個の信号を時間軸上のｍ個のサンプル（ｔ1～ｔm）に変換する。そして、ガード区間挿入器２１は、まず、ガード区間Ｔgにおいて、「ｔm-2」「ｔm-1」「ｔm」を順番に読み出して出力し、それに続く信号時間Ｔs において「ｔ1」～「ｔm」を順番に読み出して出力していく。これにより、ガード区間が挿入された信号系列が作成される。

上記構成においてガード区間の長さは、「信号時間Ｔｓ の前に出力するサンプルの数」を変えることにより制御される。この場合、ガード区間Ｔg、信号時間Ｔs、および逆フーリエ変換の周期（すなわち、シンボル周期Ｔ）が所定の関係（Ｔ＝Ｔg＋Ｔs）を満たすように、サンプル値の読出し間隔が決定される。

一例を示す。ここでは、シンボル周期＝Ｔ、ガード区間Ｔg＝０．２Ｔ、信号時間Ｔs＝０．８Ｔ、副搬送波の多重数m＝１０００であるものとする。この場合、ガード区間挿入器２１には、シンボル周期ごとに、時間軸上の１０００個のサンプル（ｔ1～ｔ1000）が入力される。そして、まず、２５０（＝１０００×０．２÷０．８）個のサンプル（ｔ751～ｔ1000）を読み出して出力する。続いて、上記１０００個のサンプル（ｔ1～ｔ1000）を読み出して出力する。このとき、サンプル値の読出し間隔は、「Ｔ／１２５０」である。また、ガード区間Ｔg＝０．１Ｔ、信号時間Ｔs＝０．９Ｔ、副搬送波の多重数m＝１０００であるものとすると、ガード区間挿入器２１は、まず、１１１（＝１０００×０．１÷０．９）個のサンプル（ｔ890～ｔ1000）を読み出して出力し、それ続いて、上記１０００個のサンプル（ｔ1～ｔ1000）を読み出して出力する。このとき、サンプル値の読出し間隔は、「Ｔ／１１１１」である。

なお、実施形態では、複数の副搬送波が合成された後にガード区間が挿入されているが、原理的には、副搬送波ごとにガード区間を挿入することも可能である。

図１４は、受信装置において受信波からガード期間を削除する処理を実現する構成の実施例である。ここでは、図１１～図１３に示すようにして作成された信号列（ｔm-2, ｔm-1, ｔm, ｔ1, ｔ2, ｔ3,... ｔm）が受信されるものとする。なお、図７に示す受信装置では、副搬送波変調を行った後にガード区間が削除されるように描かれているが、図１４に示す構成では、これらの処理は一体的に実行される。

ガード区間削除器３１は、スイッチ４１およびシフトレジスタ４２を備える。そして、信号系列（tm-2，tm-1，tm，t1，t2，t3,... tm）を受信すると、スイッチ４１を適切にＯＮ／ＯＦＦ制御することにより、ガード区間に配置されている所定数のサンプル値（ここでは、tm-2，tm-1，tm）を廃棄し、後続のm個のサンプル値（t1～tm）をシフトレジスタ４２に送る。ここで、ガード区間削除器３１は、送信装置において挿入されたガード区間の長さ（あるいは、ガード区間内のサンプル数）を認識しており、それに基づいてスイッチ４１のＯＮ／ＯＦＦ状態を制御する。一方、副搬送波復調器１３として動作するフーリエ変換器は、シフトレジスタ４２にm個のサンプル値が蓄積されると、それらのサンプル値についてフーリエ変換を行うことにより、副搬送波ごとの信号f1～fmを得る。なお、この処理は、シンボル周期Ｔごとに繰り返し実行される。

このように、実施形態のセルラ通信システムでは、送信装置（基地局）から受信装置（移動機）へ信号を送信する際、それらの間の最大伝送遅延差に基づいて、ガード区間の長さ、および送信電力が決定される。ここで、送信装置と受信装置との間の距離が短い場合は、最大伝送遅延差が小さくなり、ガード区間が短くなる。そして、ガード区間が短くなると、それに応じて受信装置において信号再生に寄与する信号時間が長くなるので、送信電力を低くすることができる。したがって、システム全体として干渉電力が減少し、伝送容量が増加することになる。

次に、上述の送信装置および受信装置の実施例を説明する。

第1の実施例：

図１５および図１６は、第１の実施例の送信装置および受信装置の構成図である。これらの装置の基本構成は、それぞれ、図６に示した送信装置および図７に示した受信装置と同じである。ただし、第１の実施形態の送信装置は、時

間多重された複数の信号系列を１つのＯＦＤＭ－ＣＤＭユニット（拡散変調器１、副搬送波変調器２、加算器３、ガード区間挿入器２１）により一括して変調することができる。

すなわち、信号系列Ｓi1 および信号系列Ｓi2 は、図１７に示すように、時間多重化部（TDMi）５１により多重化される。ここでは、これらの信号系列は、互いに異なる最大伝送遅延差を有する回線を介して伝送されるものとする。そして、この信号系列は、拡散変調器１および副搬送波変調器２により変調された後、ガード区間挿入器２１に与えられる。

ガード区間挿入器２１は、入力される信号系列に対して、対応する最大伝送遅延差よりも広いガード区間を挿入する。ここで、各信号系列に対するガード区間は、ガード区間制御部２３により設定される。また、利得調整器２２は、挿入されたガード区間に応じて決まる利得係数αを送信信号に乗算する。具体的には、図１７に示す例では、信号系列Ｓi1 が入力されている期間は、シンボル周期ごとにガード区間Ｔg1 が挿入され、信号の送信電力が「Ｐ1」になるように利得係数$\alpha i(t)$が制御される。一方、信号系列Ｓi2 が入力されている期間は、シンボル周期ごとにガード区間Ｔg2 が挿入され、信号の送信電力が「Ｐ2」になるように利得係数$\alpha i(t)$が制御される。

そして、上述のようにして変調された信号は、他の系の信号と合成された後、アンテナ７を介して送信される。

受信装置の基本的な動作は、図７を参照しながら説明した通りである。ただし、この受信装置は、自分宛ての信号のみを再生する。例えば、信号系列Ｓi1 および信号系列Ｓi2 が時間多重された信号から信号系列Ｓi1 を再生する場合には、ガード区間制御部３２は、信号系列Ｓi1 を受信している期間に、ガード区間Tg1 を削除するようにガード期間削除器３１に対して指示を与える。そして、ガード区間削除器３１は、その指示に従って信号系列Ｓi1 のシンボル周期ごと

にガード区間を削除する。このとき、信号系列Si2 を受信している期間は、ガード区間は削除される必要はない。

ガード区間削除器３１の出力は、拡散復調器１５により逆拡散復調される。このとき、拡散復調器１５は、ガード区間Tg1 が削除された信号時間Ts1 について逆拡散復調を行う。そして、分離部（DML）５２は、復調された信号から、信号系列Si1 に対応する時間スロットにおいてデータを出力する。

このように、第１の実施例の通信システムでは、時間多重された複数の信号系列を１つのＯＦＤＭ－ＣＤＭユニット（拡散変調器１、副搬送波変調器２、加算器３、ガード区間挿入器２１）により一括して変調できる。

第２の実施例：

第２の実施例の通信システムは、第１の実施例の通信システムの変形例である。すなわち、第１の実施例のシステムでは、時間多重された信号系列Si1 および信号系列Si2 がＯＦＤＭ－ＣＤＭを利用して伝送される。ここで、信号系列Si1 および信号系列Si2 は、基本的に、それぞれ対応する移動機に送信されることを想定している。これに対して、第２の実施例のシステムでは、時間多重された報知情報Bi および信号系列Si1 がＯＦＤＭ－ＣＤＭを利用して伝送される。ここで、信号系列Si1 は、所定の１または複数の受信装置に対して送信されるが、報知情報Biは、サービスエリア内のすべての受信装置（移動機）に対して送信される。したがって、この報知情報Biは、サービスエリア内の最も遠くに位置する受信装置（すなわち、最大伝送遅延差が最も大きくなる受信装置）に適切に伝送されるようなガード区間が設定され、且つ、送信電力が決定される必要がある。

図１８および図１９は、第２の実施例の送信装置および受信装置の構成図である。これらの装置の基本構成は、それぞれ、図１５に示した送信装置および図１６に示した受信装置と同じである。

第２の実施例では、ガード区間挿入器２１は、図２０に示すように、ガード区間制御部２３からの指示に従って、報知情報Ｂｉが入力されている期間は、シンボル周期ごとにガード区間Ｔg1を挿入し、信号系列Ｓi1が入力されている期間は、シンボル周期ごとにガード区間Ｔg2 を挿入する。ここで、報知情報Ｂｉに対して挿入されるガード区間Ｔg1 は、サービスエリア内において生じる最も大きな最大伝送遅延差よりも長くなるように設定される。例えば、図５において、基地局から移動機ＭＳ１～ＭＳ３へ報知情報を送信する際、基地局から移動機ＭＳ３への回線の最大伝送遅延差が最も大きかったとすると、ガード区間Ｔg1 の長さは、その最大伝送遅延差よりも長くなるように設定される。一方、信号系列Ｓi1 に対して挿入されるガード区間Ｔg2 は、対応する受信装置への回線の最大伝送遅延差よりも長くなるように設定される。例えば、図５において、基地局から移動機ＭＳ１信号系列Ｓi1 を送信する際には、ガード区間Ｔg2 の長さは、基地局から移動機ＭＳ１への回線の最大伝送遅延差よりも長くなるように設定される。

また、利得調整器２２は、ガード区間挿入器２１により挿入されたガード区間に応じた利得係数αを送信信号に乗算する。具体的には、図２０に示す例では、利得係数αi(t)は、報知情報Ｂｉを伝送するための信号の送信電力が「Ｐ1」となり、信号系列Ｓi1 を伝送するための信号の送信電力が「Ｐ2」になるように制御される。したがって、このように制御される利得係数αを送信信号に乗算することにより、報知情報Ｂｉはサービスエリア内のすべての受信装置に伝送されるように大きな送信電力で送信され、信号系列Ｓi1 は対応する受信装置に伝送される範囲で必要最小限の送信電力で送信される。

受信装置では、ガード区間制御部３２は、報知情報Ｂｉを受信している期間はガード区間Ｔg1 を指示し、信号系列Ｓi1 を受信している期間はガード区間Ｔg2 を指示する。そして、ガード区間削除器３１は、ガード区間制御部３２からの

指示に従って受信信号からガード区間を削除する。さらに、ガード区間が削除された信号は、拡散復調器１５により逆拡散された後、分離部５２により報知情報Biおよび信号系列Si1に分離される。

なお、報知情報Biに対して挿入されるガード区間Tg1の長さは、例えば、以下のようにして決定される。

（１）通信エリアの大きさに基づいて決定する。すなわち、送信装置がカバーする通信エリアの大きさに基づいて、報知情報Biが最も遅延して到着する受信装置までの遅延時間を推定し、その遅延時間に従ってガード区間Tg1の長さを決定する。

（２）報知情報Bi を送信する際の送信装置の送信電力に基づいて決定する。すなわち、報知情報Bi の送信電力により、その報知情報Bi を複数の受信装置に送信する際の伝送遅延時間の最大値を推定し、その遅延時間に従ってガード区間Tg1の長さを決定する。

（３）通信エリア内に存在する複数の受信装置との間の通信環境に基づいて決定する。すなわち、送信装置がカバーする通信エリア内に存在する複数の受信装置との間の通信環境をそれぞれ求め、これに基づいてガード区間Tg1の長さを決定する。具体的には、通信環境が最も厳しい受信装置に合わせてガード区間Tg1の長さを決定する。

（４）通信エリア内の最大遅延時間に基づいて決定する。すなわち、送信装置から通信エリア内に存在する複数の受信装置へ報知情報Bi を送信したときの遅延時間を受信装置ごとに測定し、それらのうちの最大遅延時間に基づいてガード区間Tg1の長さを決定する。

第３の実施例：

第３の実施例の通信システムでは、送信装置から受信装置へ信号が伝送されたときの最大伝送遅延差を検出し、その検出結果に基づいてガード区間および

送信電力が決定される。したがって、第３の実施例における送信装置および受信装置は、そのための機能を備えている。

図２１は、第３の実施例の送信装置の構成図である。この送信装置は、対応する受信装置において検出された最大伝送遅延差を表す最大伝送遅延差情報（τ）を受け取り、その情報に基づいてガード区間および送信電力を決定する機能を備えている。即ち、ガード区間制御部（ＧＩＮＳＣＮＴ）６１は、対応する受信装置において検出された最大伝送遅延差に基づいて、挿入すべきガード区間の長さを決定する。具体的には、ガード区間制御部６１ｉは、信号系列Si1および／または信号系列Si2を受信する受信装置から送られてくる最大伝送遅延差情報（τi）に基づいて、信号系列Si1および／または信号系列Si2を伝送するための信号に挿入すべきガード区間を決定する。また、電力制御部（ＰＣＮＴ）６２は、対応する受信装置において検出された最大伝送遅延差に基づいて、利得係数αを決定する。具体的には、電力制御部６２ｉは、信号系列Si1および／または信号系列Si2を受信する受信装置から送られてくる最大伝送遅延差情報（τi）に基づいて、信号系列Si1および／または信号系列Si2を伝送するための信号に乗算すべき利得係数αを決定する。

そして、ガード区間挿入器２１は、シンボル周期ごとに、送信信号に対してガード区間制御部６１により決定されたガード区間を挿入する。また、利得調整器２２は、電力制御部６２により決定された利得係数αを送信信号に乗算することにより、ガード区間の長さに対応する送信電力を実現する。

図２２は、第３の実施例の受信装置の構成図である。この受信装置は、送信装置から送られてきた信号の最大伝送遅延差を検出する機能を備えている。すなわち、遅延差検出部（ＤＭＥＳ）６３は、受信したベースバンド信号Srxから最大伝送遅延差を検出し、その検出結果を表す最大伝送遅延情報をガード区間制御部６４および対応する送信装置に通知する。ガード区間制御部６４は、

遅延差検出部６３からの通知に従ってガード区間を決定し、それをガード区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図２３は、図２２に示す遅延検出部６３の一例の構成図である。遅延差検出部６３は、ベースバンド信号Srx を時間Ts だけ遅延させる遅延回路７１、乗算器７２ａおよび積分器７２ｂから構成される相関検出回路７２、相関検出回路７２により検出された相関値と予め決められている所定のしきい値とを比較する比較回路７３、および比較回路７３による比較結果に基づいて最大伝送遅延差を検出する検出回路７４を含む。ここで、乗算器７２ａは、ベースバンド信号Srx にその遅延信号を乗算し、積分器７２ｂは、乗算器７２ａの出力を積分する。以下、図２４を参照しながら遅延差検出部６３の動作を説明する。

相関検出回路７２には、ベースバンド信号Srx およびそのベースバンド信号Srx を時間Ts だけ遅延させた信号（遅延信号）が入力される。ここで、各シンボル周期内のガード区間Tg には、図１１～図１３を参照しながら説明したように、信号時間Ts の最後尾部分のサンプル値が複写されている。このため、ベースバンド信号Srx とその遅延信号との間では、ベースバンド信号Srx の最後尾部分と遅延信号のガード区間とが重なったときに相関（自己相関）が高くなる。ただし、送信装置と受信装置との間に伝送遅延の異なる複数のパスが存在する場合には、各パスを介して信号を受信するごとに相関値のピークが発生する。したがって、比較回路７３を用いて上記相関値と予め設定されているしきい値とを比較すれば、各パスを介して信号を受信したタイミングをそれぞれ検出できる。よって、最初に信号を受信したタイミングと、最後に信号を受信したタイミングとの時間差を測定することにより、最大伝送遅延差が検出される。例えば、図４に示す通信環境においては、図２５に示すようにして最大伝送遅延差が検出される。

このように、第３の実施例では、送信装置と受信装置との間の回線の最大伝送遅延差が測定され、その結果に基づいてガード区間が挿入／削除されるので、ガード区間の幅を動的に変化させることが可能である。また、上記最大伝送遅延差の測定結果に従って送信信号の利得係数が決定されるので、常に、送信電力を必要最小限に抑えられる。

第４の実施例：

第４の実施例の通信システムでは、送信装置と受信装置との間の伝送距離を推定し、その推定結果に基づいてガード区間および送信電力が決定される。したがって、第４の実施例における送信装置および受信装置は、そのための機能を備えている。

図２６は、第４の実施例の送信装置の構成図である。この送信装置は、対応する受信装置との間の伝送距離の推定値を表す伝送距離情報（Ｌ）を受け取り、その情報に基づいてガード区間および送信電力を決定する機能を備えている。即ち、ガード区間制御部（ＧＩＮＳＣＮＴ）８１は、送信装置と受信装置との間の伝送距離に基づいて、挿入すべきガード区間の長さを決定する。具体的には、ガード区間制御部８１ｉは、信号系列Si1 および／または信号系列Si2 を受信する受信装置から送られてくる伝送距離情報（Li）に基づいて、信号系列Si1 および／または信号系列Si2 を伝送するための信号に挿入すべきガード区間を決定する。また、電力制御部（ＰＣＮＴ）８２は、上記伝送距離に基づいて、利得係数αを決定する。具体的には、電力調整部８２ｉは、信号系列Si1 および／または信号系列Si2 を受信する受信装置から送られてくる伝送距離情報（Li）に基づいて、信号系列Si1 および／または信号系列Si2 を伝送するための信号に乗算すべき利得係数αを決定する。

そして、ガード区間挿入器２１は、シンボル周期ごとに、送信信号に対してガード区間制御部８１により決定されたガード区間を挿入する。また、利得調

整器２２は、電力制御部８２により決定された利得係数αを送信信号に乗算することにより、ガード区間の長さに対応する送信電力を実現する。

図２７は、第４の実施例の受信装置の構成図である。この受信装置は、送信装置と当該受信装置との間の伝送距離を推定する機能を備えている。すなわち、距離推定部（ＬＭＥＳ）８３は、受信したベースバンド信号Ｓrx に基づいて送信装置と当該受信装置との間の伝送距離を推定し、その推定結果を表す伝送距離情報Ｌをガード区間制御部８４および対応する送信装置に通知する。ガード区間制御部８４は、距離推定部８３からの通知に従ってガード区間を決定し、それをガード区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図２８は、図２７に示す距離推定部８３の一例の構成図である。距離推定部８３は、第３の実施例において説明した遅延差検出部６３および変換テーブル８５から構成される。

送信装置と受信装置との間の伝送距離は、その間の回線の最大伝送遅延差と相関があり、伝送距離が長くなるほど最大伝送遅延差も大きくなることが知られている。したがって、これらの間の関係を実験またはシミュレーション等により予め求めておけば、最大伝送遅延差を検出することによって伝送距離を推定することができる。このため、距離推定部８３の変換テーブル８５には、伝送距離と最大伝送遅延差との関係を表す情報が格納されている。そして、遅延差検出部６３により検出された最大伝送遅延差をキーとしてその変換テーブル８５を検索することにより、送信装置と受信装置との間の伝送距離が推定される。

第５の実施例：

第５の実施例の通信システムでは、第４の実施例と同様に、送信装置と受信装置との間の伝送距離を推定し、その推定結果に基づいてガード区間および送

信電力が決定される。ただし、第５の実施例における推定方法は、第４の実施例のそれと異なっている。

図２９は、第５の実施例の送信装置の構成図である。この送信装置は、対応する受信装置からタイミング情報（T）を受け取ってそれに基づいて送信装置と受信装置との間の伝送距離を推定する機能、およびその伝送距離の推定値に基づいてガード区間および送信電力を決定する機能を備えている。

ガード区間制御部（ＧＩＮＳＣＮＴ）９１または電力制御部（ＰＣＮＴ）９２は、対応する受信装置から送られてくるタイミング信号Ｔに基づいて、当該送信装置と対応する受信装置との間の距離を推定する。すなわち、第５の実施例では、送信装置から信号が送信され、その信号が対応する受信装置により検出され、さらにその受信装置において上記信号が検出された旨が送信装置に通知される。ここで、上記信号が上記受信装置において検出されたタイミングは、タイミング情報Ｔを用いて送信装置に通知される。したがって、ガード区間制御部９１または電力制御部９２は、信号を送信したときから、対応する受信装置からタイミング情報Ｔを受信するまでの時間をモニタすることにより、送信装置と受信装置との間の伝送時間および伝送距離を推定できる。そして、上記伝送距離の推定値は、伝送距離情報Ｌを利用して対応する受信装置に送られる。

なお、ガード区間制御部９１は、伝送距離の推定値に基づいてガード区間の長さを決定する。また、電力制御部９２は、伝送距離の推定値に基づいて利得係数αを決定する。これらの処理は、基本的に、第４の実施例と同じである。

図３０は、第５の実施例の受信装置の構成図である。この受信装置は、送信装置から送出された信号の受信タイミングを検出する機能を備えている。すなわち、タイミング生成部（ＴＧＥＮ）９３は、受信したベースバンド信号S_{rx}を基準として受信タイミングを検出し、タイミング信号Ｔを生成する。そして、生成したタイミング信号Ｔは、送信装置へ送られる。また、ガード区間制御部

（ＧＣＮＴ）９４は、送信装置から送られてくる伝送距離情報Ｌに基づいてガード区間を決定し、それをガード区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図３１は、図３０に示すタイミング生成部９３の一例の構成図である。タイミング生成部９３は、第３の実施例において説明した遅延回路７１、相関検出回路７２、および最大値判定回路９５を含む。

上述したように、受信信号とその遅延信号との自己相関をモニタした場合、ガード区間を受信している期間の相関値が高くなる。したがって、その相関値をモニタすることにより、ガード区間の位置を検出できる。具体的には、最大値判定部９５を用いてシンボル周期ごとに上記相関値の最大値を検出することにより、ガード区間のタイミング（または、ガード区間の直後に相当するタイミング）を検出できる。そして、タイミング生成部９３は、検出したタイミングを表すタイミング情報Ｔを生成し、それを送信装置へ送る。

第６の実施例：

第６の実施例の通信システムでは、第４または第５の実施例と同様に、送信装置と受信装置との間の伝送距離を推定し、その推定結果に基づいてガード区間および送信電力が決定される。ただし、第６の実施例における推定方法は、第４または第５の実施例のそれと異なっている。

第６の実施例の通信システムでは、信号系列Ｓi1および信号系列Ｓi2を送信する際に、それらの系列にそれぞれ既知情報ＳＷが時間多重される。一方、受信装置は、受信信号に中に含まれている既知情報ＳＷを検出すると、その検出タイミングを送信装置に通知する。そして、送信装置は、既知情報ＳＷを送信したタイミングおよび対応する受信装置から送られてくるタイミング情報に基づいて、当該送信装置と受信装置との間の信号の伝送時間を検出し、その伝送時間から伝送距離を推定する。

図３２は、第６の実施例の送信装置の構成図である。この送信装置は、送信信号系列に既知情報ＳＷを多重化する機能、対応する受信装置からタイミング情報（Ｔ）を受け取ってそれに基づいて送信装置と受信装置との間の伝送距離を推定する機能、およびその伝送距離の推定値に基づいてガード区間および送信電力を決定する機能を備えている。

時間多重化部（TDM）５１は、信号系列Si1、Si2を送信する際に、それらの系列にそれぞれ既知情報ＳＷを多重する。ここで、既知情報ＳＷは、特に限定されるものではないが、対応する受信装置がそのデータパターンを認識している必要がある。

ガード区間制御部（ＧＩＮＳＣＮＴ）１０１または電力制御部（ＰＣＮＴ）１０２は、対応する受信装置から送られてくるタイミング信号Ｔに基づいて、当該送信装置と対応する受信装置との間の距離を推定する。そして、この伝送距離の推定値は、伝送距離情報Ｌを利用して対応する受信装置に送られる。なお、伝送距離を推定する方法については後述する。

なお、ガード区間制御部１０１は、伝送距離の推定値に基づいてガード区間の長さを決定する。また、電力制御部１０２は、伝送距離の推定値に基づいて利得係数αを決定する。これらの処理は、基本的に、第４または第５の実施例と同じである。

図３３は、第６の実施例の受信装置の構成図である。この受信装置は、受信波から既知情報ＳＷを分離して出力する機能、および既知情報を受信した旨を送信装置に通知する機能を備えている。すなわち、タイミング生成部（ＴＧＥＮ）１０３は、分離部（ＤＭＬ）５２から出力された既知情報ＳＷを検出すると、その検出タイミングから所定時間経過後にタイミング信号Ｔを生成して送信装置へ送出する。また、ガード区間制御部（ＧＣＮＴ）１０４は、送信装置から送られてくる伝送距離情報Ｌに基づいてガード区間を決定し、それをガー

ド区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図３４は、図３３に示すタイミング生成部１０３の一例の構成図である。タイミング生成部１０３には、当該受信装置により復調された信号列が入力される。ここで、この信号列は、送信装置において挿入された既知情報ＳＷを含んでいる。そして、この信号列は、既知情報ＳＷのワード長と等しい長さのシフトレジスタ１０５に順番に入力されていく。論理反転回路１０６、加算回路１０７、および比較回路１０８は、シフトレジスタ１０５に新たなデータの書き込まれるごとに、保持されているデータが既知情報ＳＷと一致するか否かを調べる。なお、論理反転回路１０６は、既知情報ＳＷのワードパターンに対応して設けられている。また、加算回路１０７は、シフトレジスタ１０５に保持されている各エレメントの値またはシフトレジスタ１０５に保持されている各エレメントの値の論理反転値を加算する。そして、比較回路１０８は、加算回路１０７による加算結果と予め設定されている閾値とを比較し、加算結果の方が大きかったときにタイミング信号Ｔを出力する。

このように、第６の実施例の通信システムでは、送信装置から受信装置へ既知情報ＳＷが送信され、その既知情報ＳＷを検出した旨が受信装置から送信装置へ通知される。したがって、送信装置から受信装置へ信号が伝送される際の伝送時間を「Ｔ1」、受信装置が既知情報ＳＷを検出してからタイミング情報を送信するまでの時間を「Ｔd」、受信装置から送信装置へタイミング情報が伝送される際の伝送時間を「Ｔ2」、既知情報ＳＷを送信してからタイミング情報を受信するまでの時間を「Ｔ0」とすると、下記の式が成立する。なお、「Ｔ2」は「Ｔ1」に比例するものとし、その比例定数を「β」とする。

$$
\begin{aligned}
T1 &= T0 - Td - T2 \\
&= T0 - Td - \beta \cdot T1
\end{aligned}
$$

$\therefore\ T1 = (T0 - Td) / (1 + \beta)$

ここで、送信装置と受信装置との間の伝送距離は、送信装置から受信装置へ信号が伝送される際の伝送時間（Ｔ1）に比例する。また、受信装置が既知情報ＳＷを検出してからタイミング情報を送信するまでの時間（Ｔd）は既知である。したがって、送信装置は、既知情報ＳＷを送信してからタイミング情報を受信するまでの時間（Ｔ0）を測定することにより、送信装置と受信装置との間の伝送距離を推定できる。なお、この実施例では、ガード区間制御部１０１または電力制御部１０２がその伝送距離を推定する。

第７の実施例：

第７の実施例の通信システムでは、ガード区間の長さを変えながら伝送エラー率が測定され、所定の伝送品質が確保されるようにガード区間の長さ（および、送信電力）が決定される。したがって、第７の実施例における送信装置および受信装置は、そのための機能を備えている。

図３５は、第７の実施例の送信装置の構成図である。この送信装置は、既知パターンデータ（ＰＬj）を変調して送信する機能、および対応する受信装置から最大伝送遅延差情報（τ）を受け取ってそれに基づいてガード区間および送信電力を決定する機能を備えている。

既知パターンデータ（ＰＬj）は、拡散変調器１により拡散された後、副搬送波変調器２により変調される。ここで、既知パターンデータ（ＰＬj）は、特に限定されるものではないが、各受信装置により認識されているものとする。また、拡散変調器１は、既知パターンデータ（ＰＬj）に対応する拡散符号Ｃ（ＰＬj）により拡散される。

ガード区間挿入器（ＧＩＮＳj）２１は、シンボル周期ごとに、既知パターンデータ（ＰＬj）を伝送するための信号系列に比較的長いガード区間を挿入する。ここで、このガード区間は、例えば、サービスエリア内の最も遠い位置にいる

移動機（受信装置）へ信号を送信する場合を想定して決定されるようにしてもよい。また、利得調整器（Gj）２２は、ガード区間が挿入された信号系列が十分に大きな送信電力で送信されるように適切な利得係数αjを乗算する。ここで、この利得係数αjは、例えば、サービスエリア内の最も遠い位置にいる移動機（受信装置）へ信号を送信する場合を想定して決定されるようにしてもよい。そして、既知パターンデータ（ＰＬj）は、信号系列Ｓi1、Ｓi2 と合成されて送信される。

ガード区間制御部（ＧＩＮＳＣＮＴ）６１および電力制御部（ＰＣＮＴ）６２の動作は、第３の実施例において説明した通りである。すなわち、ガード区間制御部６１は、対応する受信装置から送られてくる最大伝送遅延差情報に基づいて、挿入すべきガード区間の長さを決定する。また、電力制御部６２は、対応する受信装置から送られてくる最大伝送遅延差情報に基づいて、利得係数αを決定する。

図３６は、第７の実施例の受信装置の構成図である。この受信装置は、既知パターンデータ（ＰＬj）を抽出してその伝送エラーを測定する機能、および伝送エラー率に基づいて最大伝送遅延差情報を生成する機能を備えている。

受信波は、復調回路により復調される。このとき、拡散復調器（ＳＤＥＭ）１５において、信号系列Ｓi1 を復調するときは拡散符号Ｃi が使用され、既知パターンデータ（ＰＬj）を復調するときには拡散符号Ｃ（ＰＬj）が使用される。そして、分離部５２は、再生された信号列を、信号系列Ｓi1 および既知パターンデータ（ＰＬj）に分離する。

遅延差検出部（ＤＭＥＳ）１１１は、再生された既知パターンデータ（ＰＬj）の伝送エラー率を測定し、その伝送エラー率に基づいて最大伝送遅延差情報τを生成する。この最大伝送遅延差情報τは、ガード区間制御部（ＧＣＮＴ）１１２に与えられると共に、送信装置に送られる。そして、ガード区間制御部１

１２は、その最大伝送遅延差情報τに基づいてガード区間を決定し、それをガード区間削除器３１に指示する。そして、ガード区間削除器３１が、その指示に従って受信信号からガード区間を削除する。

図３７は、図３６に示す遅延差検出部１１１の動作を示すフローチャートである。ここでは、予め複数のガード区間長データτ0〜τn が用意されているものとする。また、ガード区間長データτ0〜τn の中で、「τ0」が最小であり、「τn」が最大であるものとする。なお、このフローチャートの処理は、たとえば、既知パターンデータ（ＰＬj）を受信するごとに実行される。

ステップＳ１では、拡散復調器１５に拡散符号Ｃ（ＰＬj）を設定する。ここで、この拡散符号Ｃ（ＰＬj）は、送信装置において既知パターンデータ（ＰＬj）を拡散する際に使用されてものである。これにより、以降、受信信号が逆拡散されると、既知パターンデータ（ＰＬj）が再生されることになる。ステップＳ２では、ガード区間長データを指定するための変数を初期化する。すなわち、「ｉ＝０」が設定される。

ステップＳ３では、ガード区間制御部１１２にガード区間長データτｉを設定する。ただし、この時点では、「ｉ＝０」であるので、ガード区間制御部１２３には「ガード区間長データτ0」が設定されることになる。ここで、「ガード区間長データτ0」は、予め用意されている候補データの中で最も短い値を持っている。また、このとき、分離部５２は、再生された既知パターンデータ（ＰＬj）が遅延差検出部１１１に導かれるように出力する。

ステップＳ４では、再生された既知パターンデータ（ＰＬj）の誤り率（誤りビット数）を調べる。そして、この誤り率が予め設定されているしきい値よりも高かった場合には、十分な通信品質が得られていないものとみなし、ステップＳ５へ進む。ステップＳ５では、変数ｉをインクリメントできるか否かが調べられる。そして、可能であれば、ステップＳ６において変数ｉがインクリメ

ントされた後、ステップＳ３に戻る。

このように、ステップＳ３〜Ｓ６では、ガード区間制御部１１２に設定すべきガード区間長を少しずつ長くしていきながら、それぞれについて既知パターンデータ（ＰＬj）の誤り率が測定される。そして、既知パターンデータ（ＰＬj）の誤り率がしきい値以下になった時点で、ステップＳ７へ進む。したがって、上記処理により、所望の通信品質が得られる範囲内で、できるかぎり短いガード区間長が決定される。なお、この時点で、ガード区間制御部１１２には、最適なガード区間が設定されていることになる。

ステップＳ７では、拡散復調器１５に拡散符号Ci を設定する。ここで、拡散符号Ci は、送信装置において信号系列Si1、Si2 を拡散する際に使用されたものである。したがって、以降、拡散復調器１５は、受信信号から信号系列Si1 を復調できるようになる。ステップＳ８では、ステップＳ３〜Ｓ６において決定されたガード区間長を送信装置に通知する。

このように、第７の実施例では、伝送エラー率を測定しながら所定の通信品質が確保されるようにガード区間の長さ（および、送信電力）が決定される。したがって、必要最小限のガード区間および送信電力で所望の通信品質が確保される。

第８の実施例：

第８の実施例の通信システムは、第７の実施例の通信システムの変形例である。すなわち、第７の実施例では、受信装置に設定すべきガード区間長が決定され、その値が送信装置に通知される構成であった。これに対して、第８の実施例では、受信装置に設定すべきガード区間長に基づいて送信装置と受信装置との間の伝送距離が推定され、その推定結果が送信装置に通知される。

図３８は、第８の実施例の送信装置の構成図である。この送信装置は、基本的には、図３５に示した第７の実施例の送信装置と同じである。ただし、第８

の実施例の送信装置は、図３５に示したガード区間制御部（ＧＩＮＳＣＮＴ）６１および電力制御部（ＰＣＮＴ）６２の代わりに、ガード区間制御部（ＧＩＮＳＣＮＴ）８１および電力制御部（ＰＣＮＴ）８２が設けられている。なお、ガード区間制御部８１および電力制御部８２の動作は、第４の実施例において説明した通りである。すなわち、ガード区間制御部８１は、対応する受信装置から送られてくる伝送距離情報Ｌに基づいて、挿入すべきガード区間の長さを決定する。また、電力制御部８２は、対応する受信装置から送られてくる伝送宇距離情報Ｌに基づいて、利得係数αを決定する。

図３９は、第８の実施例の受信装置の構成図である。この受信装置は、図３６に示した第７の実施例の受信装置の遅延差検出部１１１、ガード区間制御部１１２の代わりに、距離推定部（ＬＭＥＳ）１２１、変換テーブル（ＴＢＬ）１２２、ガード区間制御部（ＧＣＮＴ）１２３を備える。ここで、距離推定部１２１およびガード区間制御部１２３は、まず、第７の実施例と同様に、最適なガード区間長を決定する。その後、距離推定部１２１は、変換テーブル１２２にアクセスし、決定したガード区間長に対応する伝送距離を取得する。そして、その伝送距離を表す伝送距離情報Ｌを送信装置に通知する。なお、変換テーブル１２２は、図２８に示した変換テーブル８５に相当し、ガード区間長と伝送距離との対応関係が格納されている。

図４０は、図３９に示す距離推定部１２１の動作を示すフローチャートである。図４０において、ステップＳ１～Ｓ７は、図３７に示した第７の実施例における処理と同じである。すなわち、ステップＳ１～Ｓ７において、受信装置に設定すべきガード区間長τiが決定される。続いて、ステップＳ１１では、変換テーブル１１２を参照して、ガード区間長τiを伝送情報Ｌｉに変換する。そして、ステップＳ１２において、ステップＳ１１で取得した伝送情報を送信装置に通知する。

このように、本発明によれば、セルラ通信システムにおける基地局とそのサービスエリア内の移動機との間の伝送路で生じる最大伝送遅延差に応じてガード区間および送信電力が適切に設定されるので、干渉の発生が低減される。あるいは、伝送路の送信帯域内での伝送容量が最適化されるので、通信システムの効率的な運用が可能となり、総伝送容量を増加させることができる。

なお、ガード区間および送信電力は、送信装置と受信装置との間の回線の最大伝送遅延差（または、伝送距離）に応じて動的に制御されてもよいし、固定的に設定されてもよい。例えば、通信の開始時にガード区間および送信電力が決定され、以降、その通信が終了するまでそれらが変化しないようにしてもよい。また、通信中に、随時、ガード区間および送信電力が動的に調整されてもよい。さらに、送信装置および受信装置の位置が変化しない場合には、初期設定処理においてガード区間および送信電力が決定されてもよい。

また、本発明では、最大伝送遅延差（または、伝送距離）に応じてガード区間および送信電力が決定されるが、ガード区間長と送信電力の関係は、たとえば、実験またはシミュレーション等により予め一意に決められていてもよい。

請求の範囲

1．　直交周波数分割多重を利用して送信装置から受信装置へ信号を伝送する通信システムであって、

上記送信装置は、

信号系列を用いて複数の副搬送波を変調する変調手段と、

上記変調手段の出力にガード区間を挿入する挿入手段と、

上記ガード区間が挿入された変調信号を送信する送信手段を有し、

上記受信手段は、

上記送信装置から送信された変調信号について副搬送波ごとにガード区間の削除処理と復調処理を行い、信号系列を再生する復調手段を有し、

上記ガード区間の長さは、上記送信装置と上記受信装置との間の通信環境に基づいて決定される通信システム。

2．　請求項1に記載の通信システムであって、

上記送信装置は、上記ガード区間の長さに応じて上記変調信号を送信する際の送信電力を制御する電力制御手段をさらに有する。

3．　直交周波数分割多重を利用して送信装置から第1の受信装置を含む複数の受信装置へ信号を伝送する通信システムであって、

上記送信装置は、

第1の受信装置へ伝送する第1の信号系列、および第1の受信装置とは異なる他の受信装置へ伝送する第2の信号系列が多重された信号系列を用いて複数の副搬送波を変調する変調手段と、

上記第1の信号系列の変調出力に第1のガード区間を、上記第2の信号系列の変調出力に第2のガード区間をそれぞれ挿入する挿入手段と、

上記第1のガード区間と第2のガード区間がそれぞれ挿入された変調信号

を送信する送信手段を有し、

上記第１の受信装置は、

上記第１のガード区間の削除処理と復調処理を行い、第１の信号系列を再生する復調手段を有し、

上記第１のガード区間の長さは、上記送信装置と上記第１の受信装置との間の通信環境に基づいて決定されるとともに、上記第２のガード区間の長さは、上記送信装置と上記他の受信装置との間の通信環境に基づいて決定される。

４．　直交周波数分割多重を利用して送信装置から第１の受信装置を含む複数の受信装置へ信号を伝送する通信システムであって、

上記送信装置は、

第１の受信装置へ伝送する第１の信号系列、および上記送信装置の通信エリア内の第１の受信装置を含む複数の受信装置に伝送する第２の信号系列が多重された信号系列を用いて複数の副搬送波を変調する変調手段と、

上記第１の信号系列の変調出力に第１のガード区間を、上記第２の信号系列の変調出力に第２のガード区間をそれぞれ挿入する挿入手段と、

上記第１のガード区間と第２のガード区間がそれぞれ挿入された変調信号を送信する送信手段を有し、

上記第１の受信装置は、

上記第１のガード区間の削除処理と第２のガード区間の削除処理と復調処理を行い、第１の信号系列と第２の信号系列を再生する復調手段を有し、

上記第１のガード区間の長さは、上記送信装置と上記第１の受信装置との間の通信環境に基づいて決定される

５．　請求項４に記載の通信システムであって、

上記第２ガード区間の長さは、通信エリア内に存在する複数の受信装置が第２の信号系列を再生できるように決定される。

６．　請求項１に記載の通信システムであって、

上記受信装置は、上記送信装置と当該受信装置との間の回線の最大伝送遅延差を検出する検出手段をさらに有し、

上記挿入手段は、上記検出手段により検出された最大伝送遅延差に基づいて決まる長さのガード区間を挿入し、

上記削除手段は、その最大伝送遅延差に従ってガード区間を削除する。

７．　請求項１に記載の通信システムであって、

上記受信装置は、上記送信装置と当該受信装置との間の伝送距離を推定する推定手段をさらに有し、

上記挿入手段は、上記推定手段により推定された伝送距離に基づいて決まる長さのガード区間を挿入し、

上記削除手段は、その推定された伝送距離に従ってガード区間を削除する。

８．　請求項１に記載の通信システムであって、

上記送信装置は、上記送信装置と当該受信装置との間の伝送距離を推定する推定手段をさらに有し、

上記挿入手段は、上記推定手段により推定された伝送距離に基づいて決まる長さのガード区間を挿入し、

上記削除手段は、その推定された伝送距離に従ってガード区間を削除する。

９．　請求項８に記載の通信システムであって、

上記推定手段は、当該送信装置から信号が送信されたときから、上記受信装置からその信号に対応する応答が返ってくるまでの時間に基づいて上記伝送距離を推定する。

１０．請求項１に記載の通信システムであって、

上記受信装置は、上記送信装置から当該受信装置へ信号が伝送されたときの通信品質をモニタするモニタ手段をさらに有し、

上記ガード区間の長さは、予め決められた所定の通信品質が満たされるように決定される。

１１．直交周波数分割多重を利用して送信装置から受信装置へ信号を伝送する方法であって、

直交周波数分割多重を利用して信号系列を変調し、

上記変調により得られた信号に対して、上記送信装置と上記受信装置との間の通信環境に基づいて決定される長さのガード区間を挿入し、

上記ガード区間が挿入された変調信号を送信する信号伝送方法。

１２．請求項１１に記載の方法であって、

上記ガード区間が挿入された変調信号は、そのガード区間の長さに応じてその送信電力が制御される。

１３．直交周波数分割多重を利用して送信装置から受信装置へ信号を伝送する方法であって、

送信相手先の異なる第１の信号系列および第２の信号系列を直交周波数多重を利用して変調し、

上記変調された第１の信号系列に対して、上記送信装置と送信相手先との間の通信環境に基づいて決定される長さの第１のガード区間を挿入し、上記変調された第２の信号系列に対して、上記送信装置と送信相手先との間の通信環境に基づいて決定される長さの第２のガード区間を挿入し、

上記第１のガード区間および第２のガード区間がそれぞれ挿入された変調信号を送信する信号伝送方法。

１４．請求項１１に記載の方法であって、

上記ガード区間の長さは、上記送信装置と上記受信装置との間の回線の最大伝送遅延差または伝送距離に基づいて決定される。

１５．請求項１１に記載の方法であって、

上記送信装置から上記受信装置へ信号が伝送されたときの通信品質をモニタし、

上記ガード区間の長さは、予め決められた所定の通信品質が満たされるように決定される。

１６． セルラ通信システムにおいて直交周波数分割多重を利用して移動機へ信号を伝送する基地局装置であって、

直交周波数分割多重を利用して信号系列を変調する変調手段と、

上記変調手段により得られた変調信号に対して、当該基地局装置と上記信号系列を送信すべき移動機との間の通信環境に基づいて決定される長さのガード区間を挿入する挿入手段と、

上記ガード区間が挿入された変調信号を送信する送信手段と、

を有する基地局装置。

１７． 請求項１６に記載の基地局装置であって、

上記ガード区間の長さに応じて上記変調信号を送信する際の送信電力を制御する電力制御手段をさらに有する。

１８． セルラ通信システムにおいて直交周波数分割多重を利用して移動機へ信号を伝送する基地局装置であって、

送信相手先の異なる第１の信号系列および第２の信号系列を直交周波数多重を利用してそれぞれ変調する変調手段と、

上記変調手段により得られた変調された第１の信号系列に対して、上記送信装置と送信相手先との間の通信環境に基づいて決定される長さの第１のガード区間を挿入するとともに、上記変調手段により得られた変調された第２の信号系列に対して、上記送信装置と送信相手先との間の通信環境に基づいて決定される長さの第２のガード区間を挿入する挿入手段と、

上記第１のガード区間および第２のガード区間がそれぞれ挿入された変調信

号を送信する送信手段と、

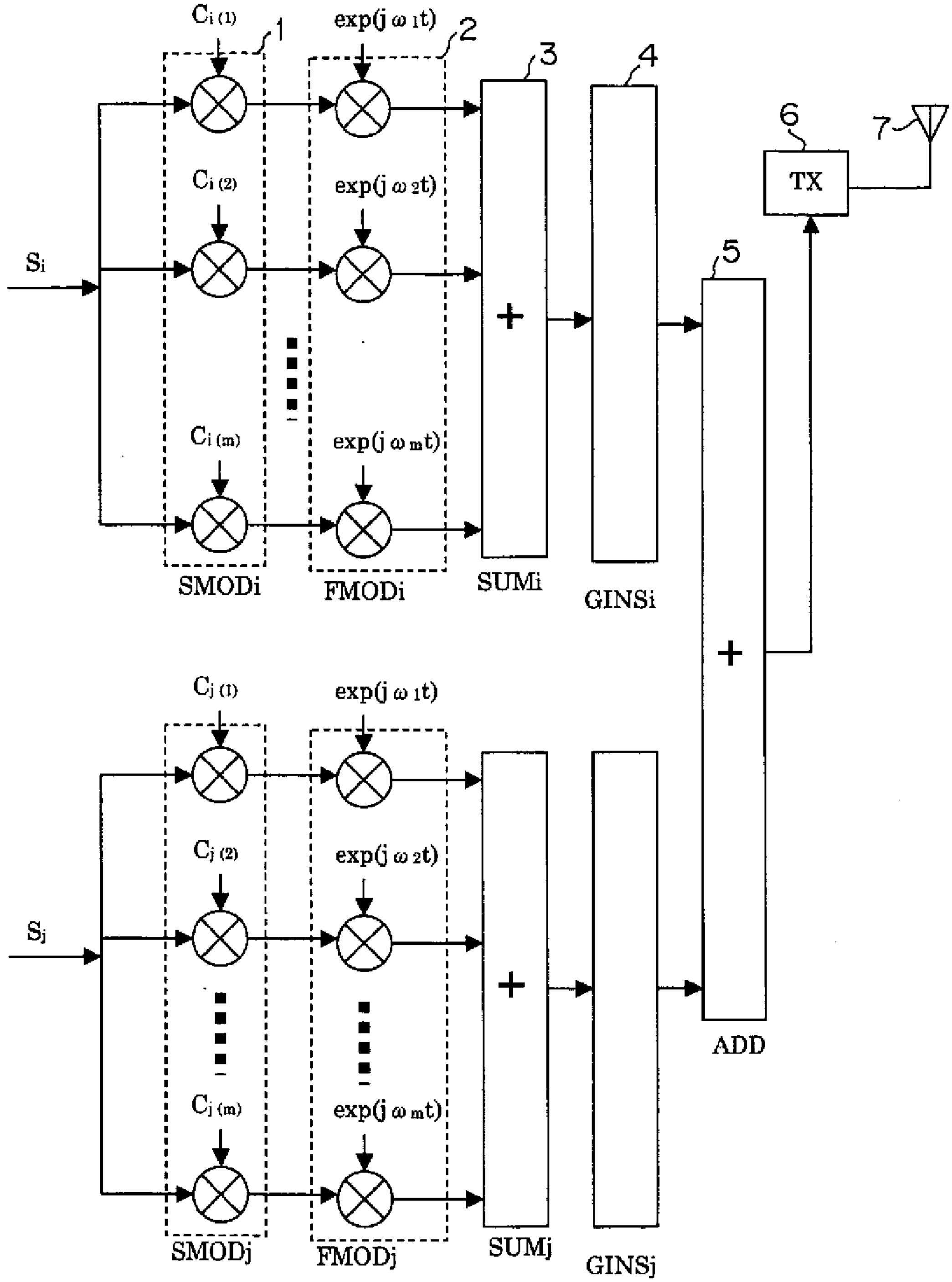
を有する基地局装置。

１９．セルラ通信システムにおいて、直交周波数分割多重を利用して基地局から送信された信号を受信する移動機であって、

受信した信号が自移動機宛ての第１の信号系列と他移動機宛ての第２の信号系列を含む場合、第１の信号系列に対応する第１のガード区間の削除処理と復調処理を行う復調手段を有する移動機。

２０．セルラ通信システムにおいて、直交周波数分割多重を利用して基地局から送信された信号を受信する移動機であって、

受信した信号が自移動機宛ての第１の信号系列と自移動機を含む複数の移動機宛ての第２の信号系列を含む場合、第１の信号系列に対応する第１のガード区間の削除処理と、第２の信号系列に対応する第２のガード区間の削除処理と、復調処理を行う復調手段を有する移動機。



Ci (1)
1
exp(j ω 1t)
2
3
4
6
7
TX
Si
Ci (2)
exp(j ω 2t)
+
5
Ci (m)
exp(j ω mt)
SMODi
FMODi
SUMi
GINSi
+
Cj (1)
exp(j ω 1t)
Sj
Cj (2)
exp(j ω 2t)
+
ADD
Cj (m)
exp(j ω mt)
SMODj
FMODj
SUMj
GINSj

図1

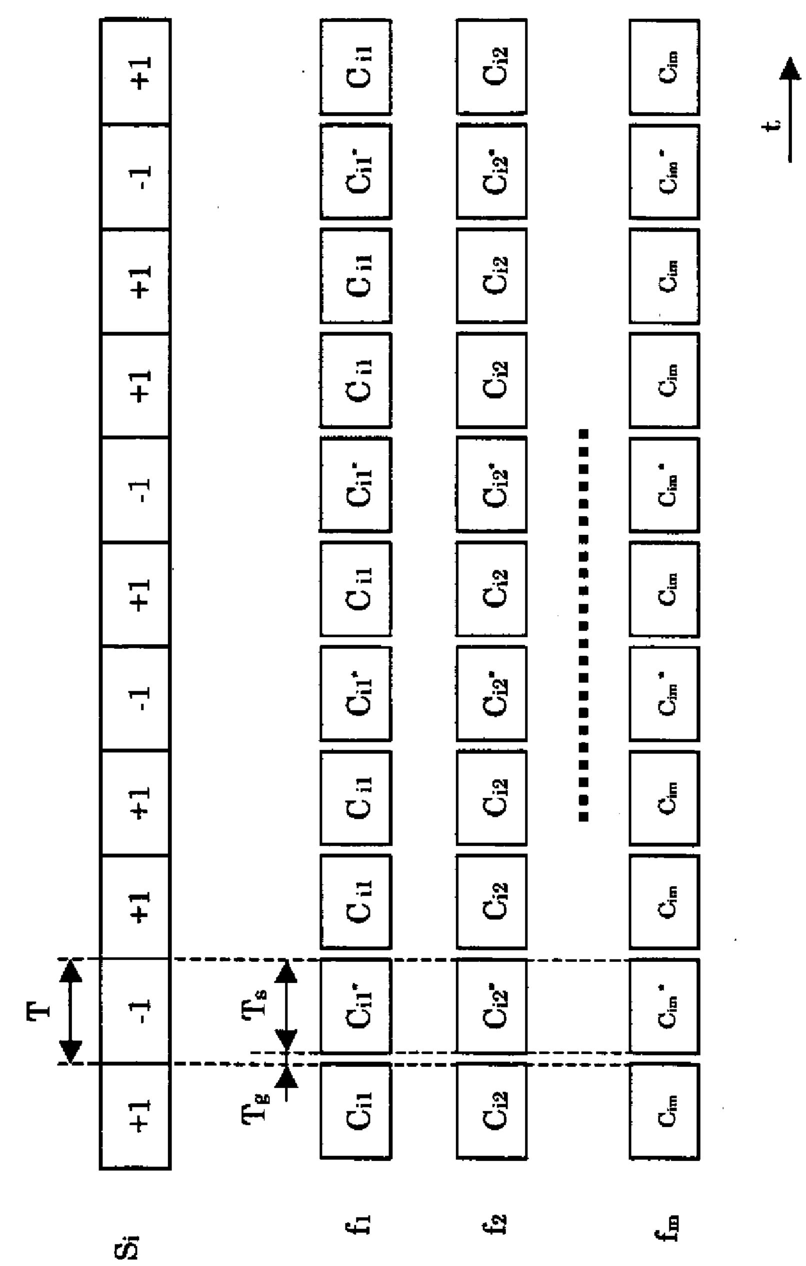


図2

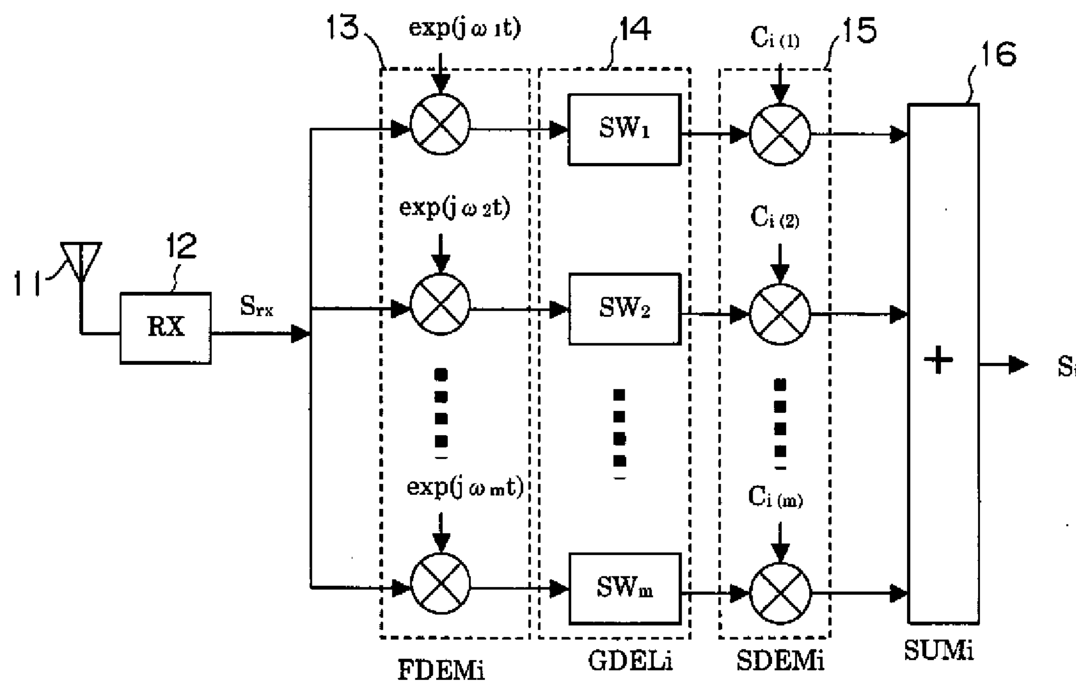

11
12
RX
Srx
13
exp(jω1t)
exp(jω2t)
exp(jωmt)
14
SW1
SW2
SWm
15
Ci(1)
Ci(2)
Ci(m)
16
+
Si
FDEMi
GDELi
SDEMi
SUMi

図3

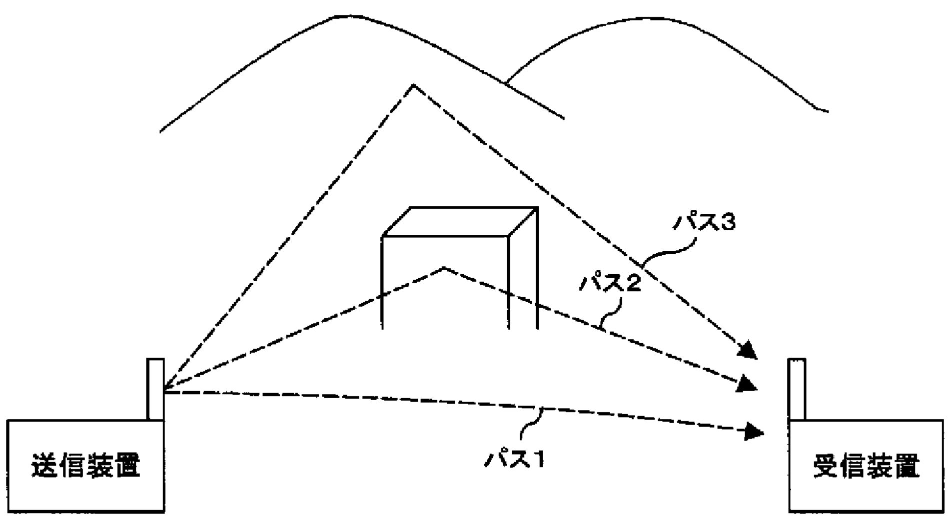


図4

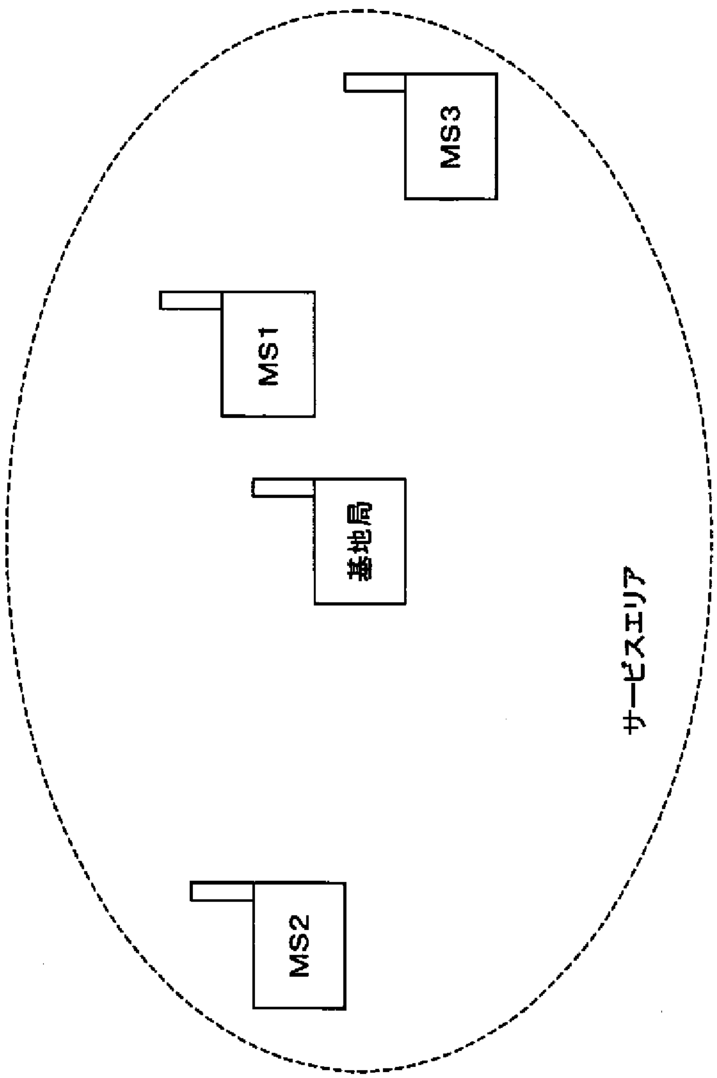


図5

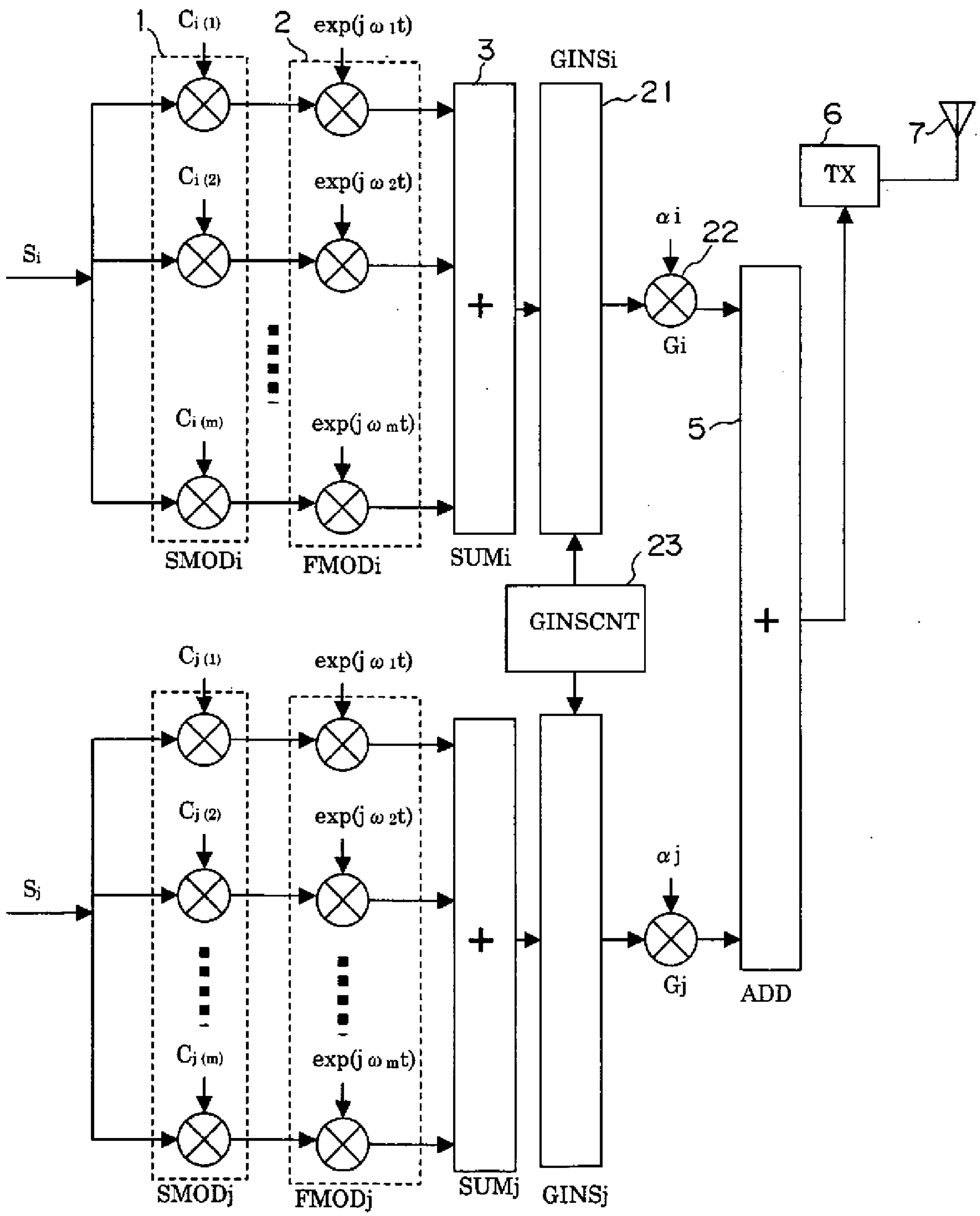

1
2
3
21
22
23
5
6
7
$C_{i(1)}$
$C_{i(2)}$
$C_{i(m)}$
$\exp(j\omega_1 t)$
$\exp(j\omega_2 t)$
$\exp(j\omega_m t)$
S_i
SMODi
FMODi
SUMi
GINSi
αi
Gi
TX
GINSCNT
$C_{j(1)}$
$C_{j(2)}$
$C_{j(m)}$
S_j
SMODj
FMODj
SUMj
GINSj
αj
Gj
ADD

図6

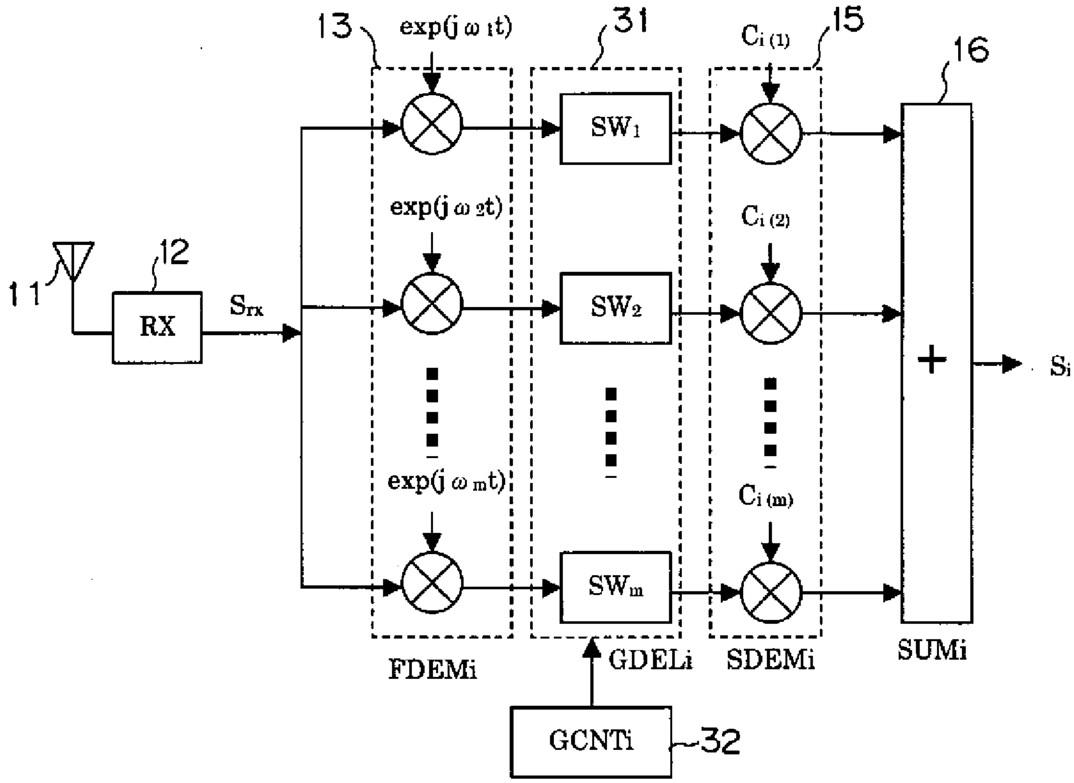

13
exp(j ω 1t)
31
Ci (1)
15
16
SW1
exp(j ω 2t)
Ci (2)
12
11
RX
Srx
SW2
+
Si
exp(j ω mt)
Ci (m)
SWm
FDEMi
GDELi
SDEMi
SUMi
GCNTi
32

図7

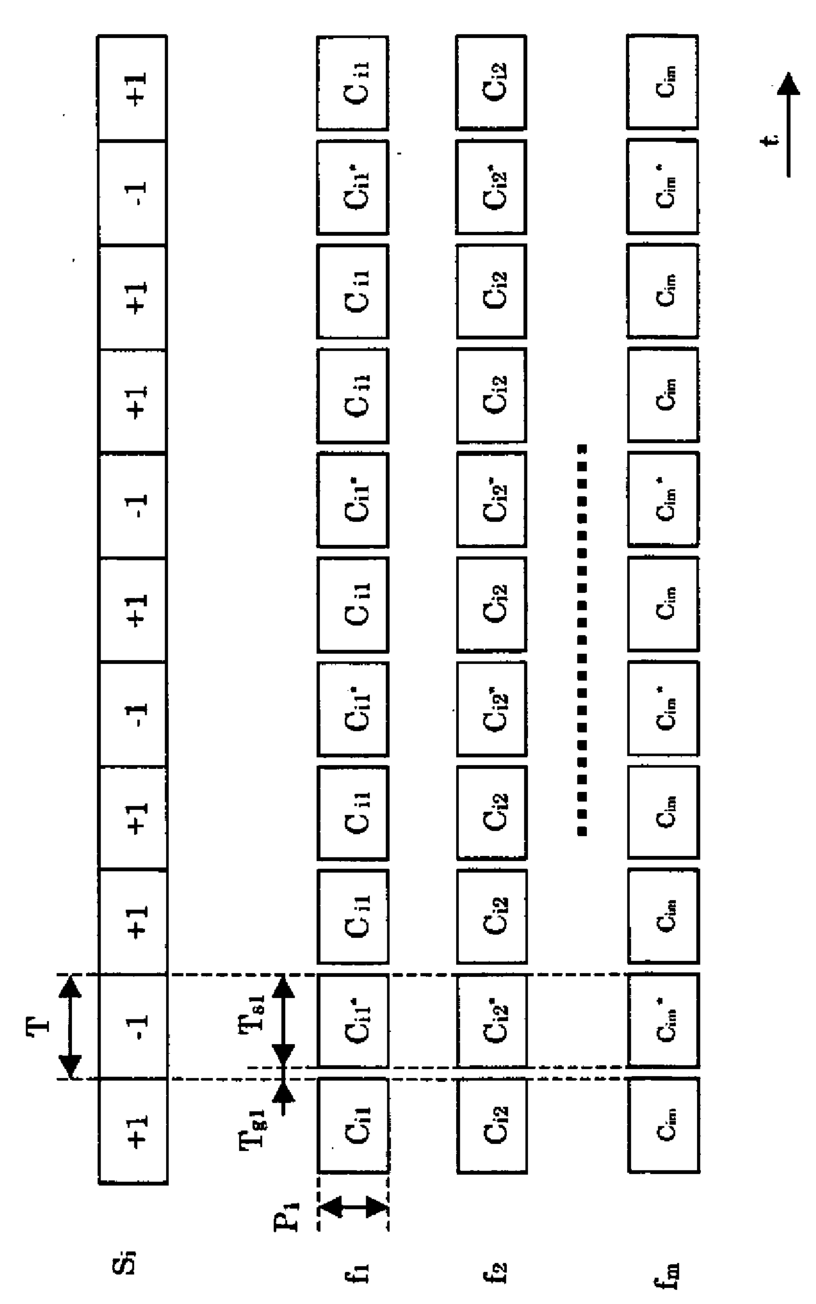


図8

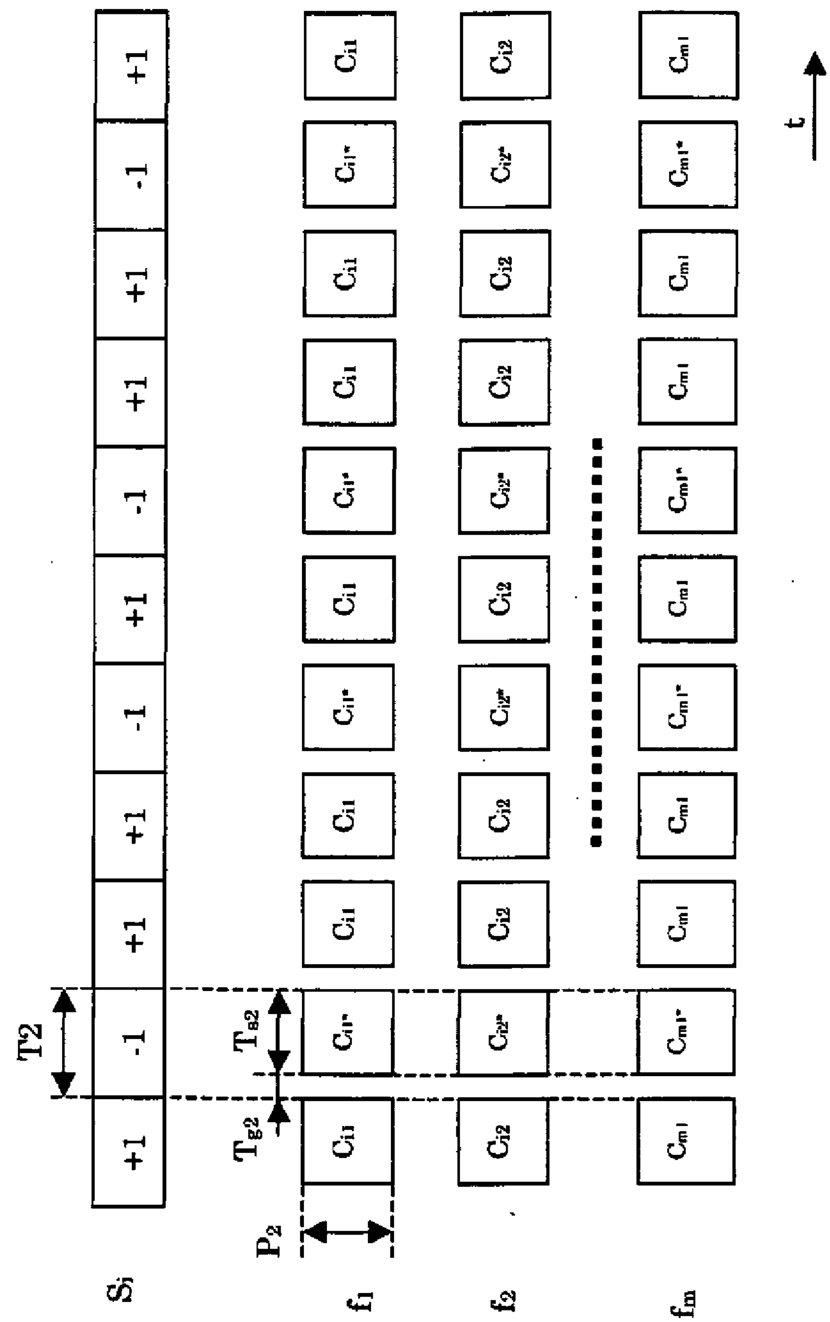


図9

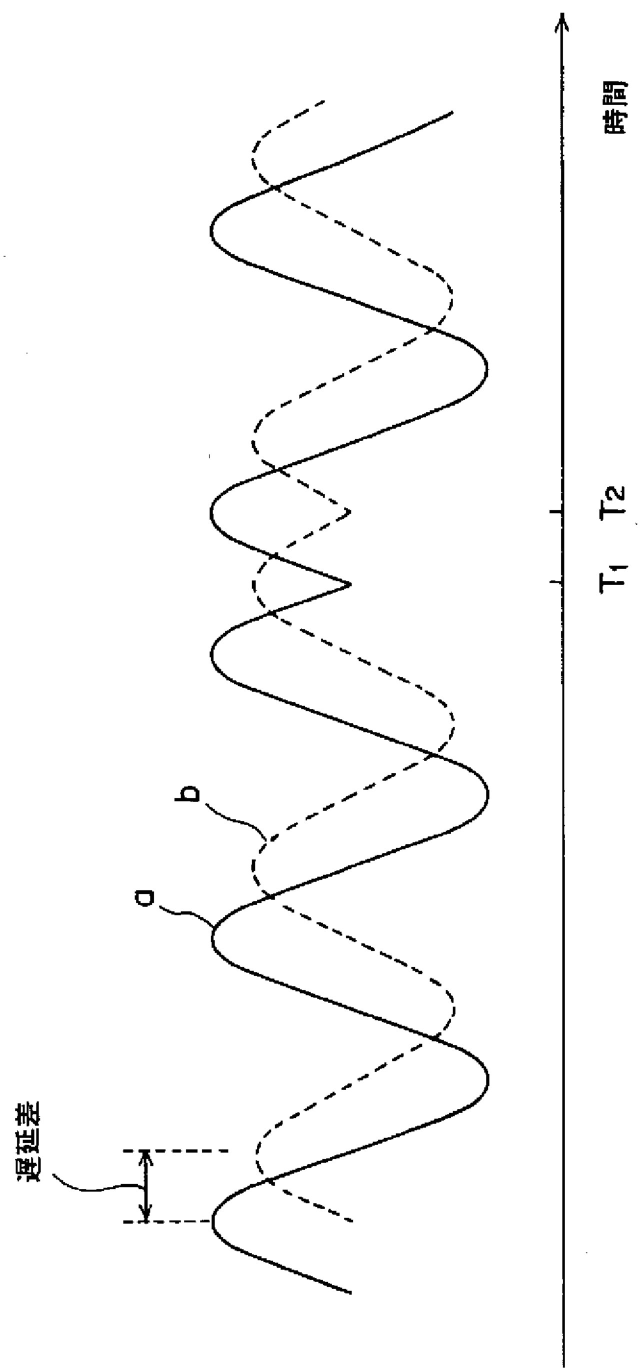


図10

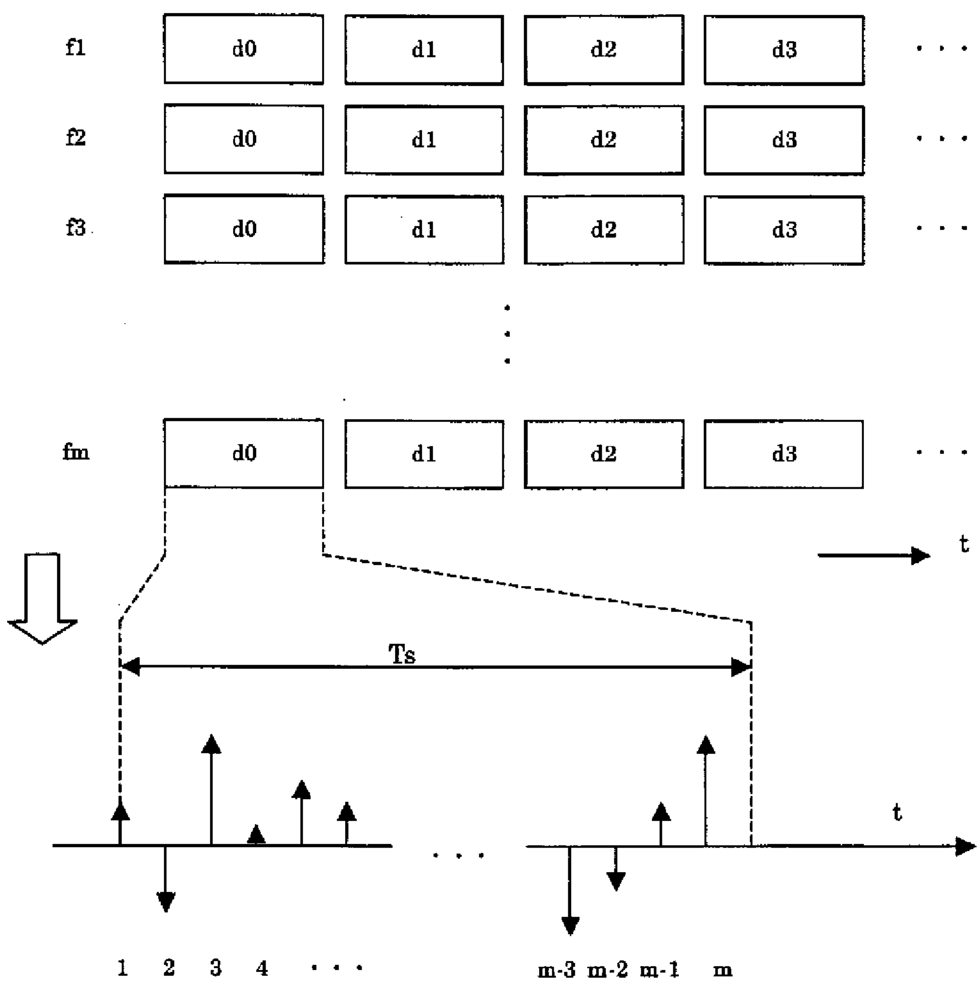


図11

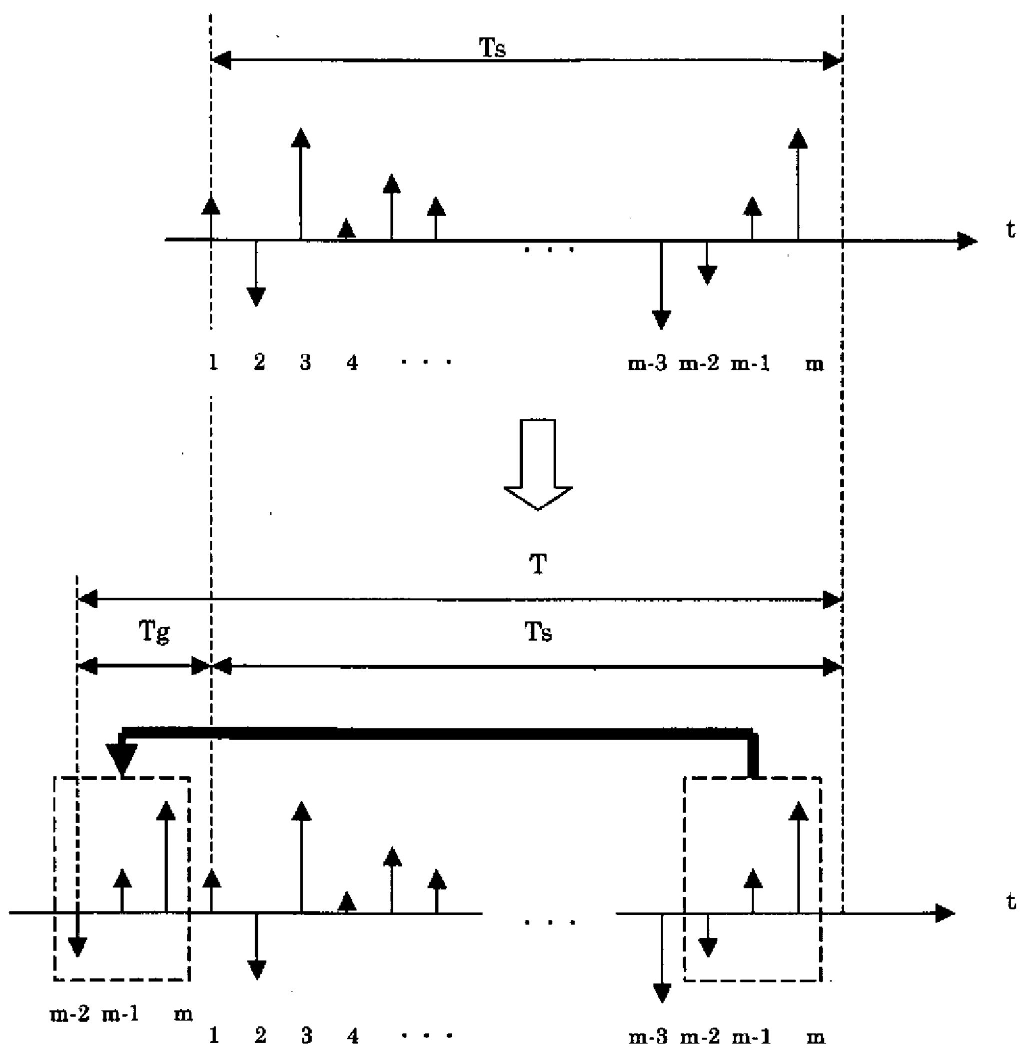


図12

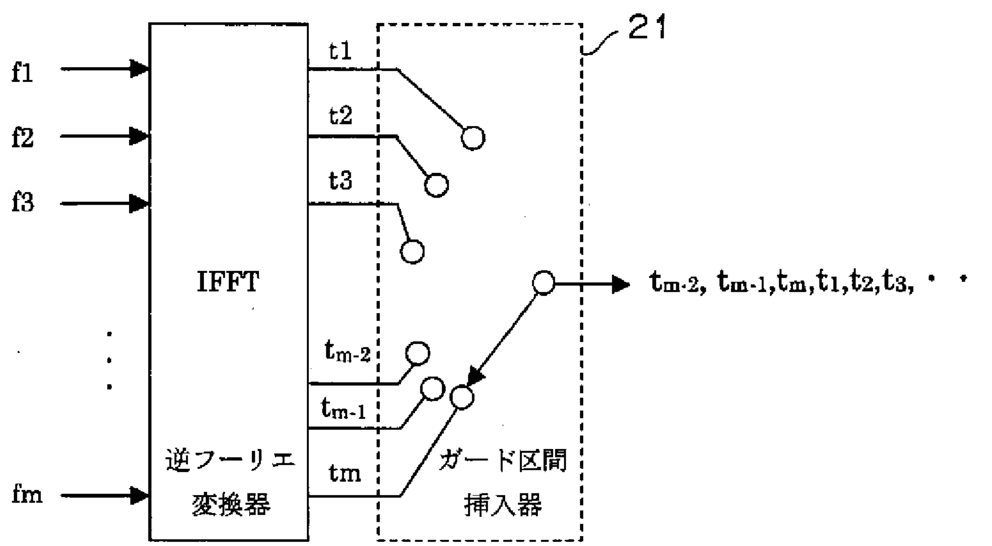
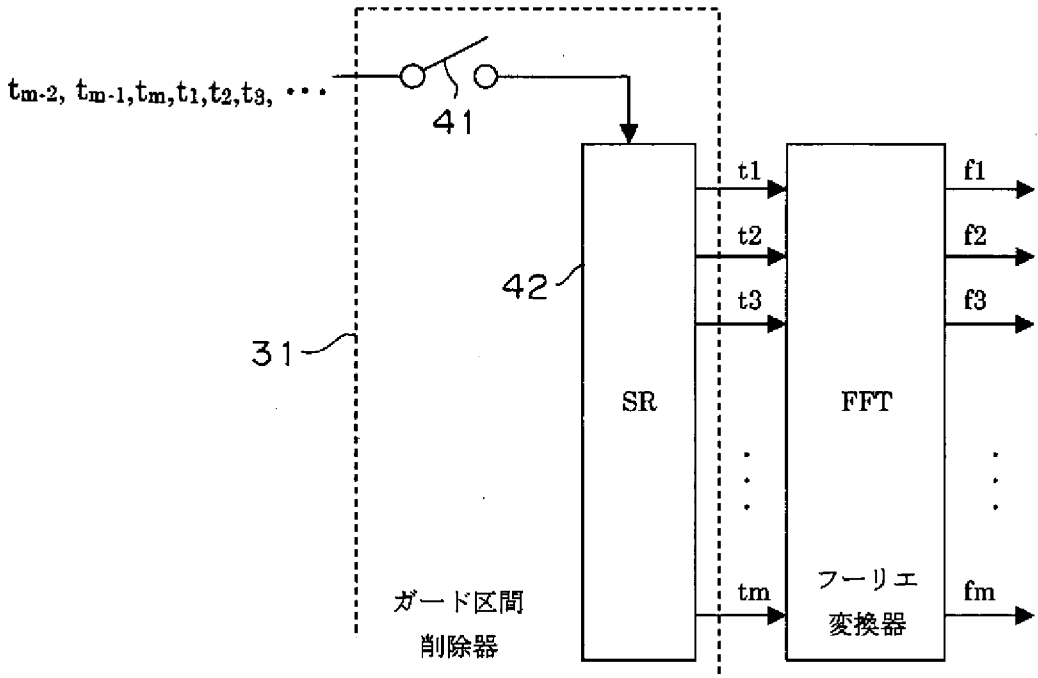

21
f1
f2
f3
fm
IFFT
逆フーリエ
変換器
t1
t2
t3
t_{m-2}
t_{m-1}
tm
ガード区間
挿入器
$t_{m-2}, t_{m-1}, t_m, t_1, t_2, t_3, \cdots$

図13



$t_{m-2}, t_{m-1}, t_m, t_1, t_2, t_3, \cdots$
41
42
31
SR
FFT
t1
t2
t3
tm
f1
f2
f3
fm
ガード区間
削除器
フーリエ
変換器

図14

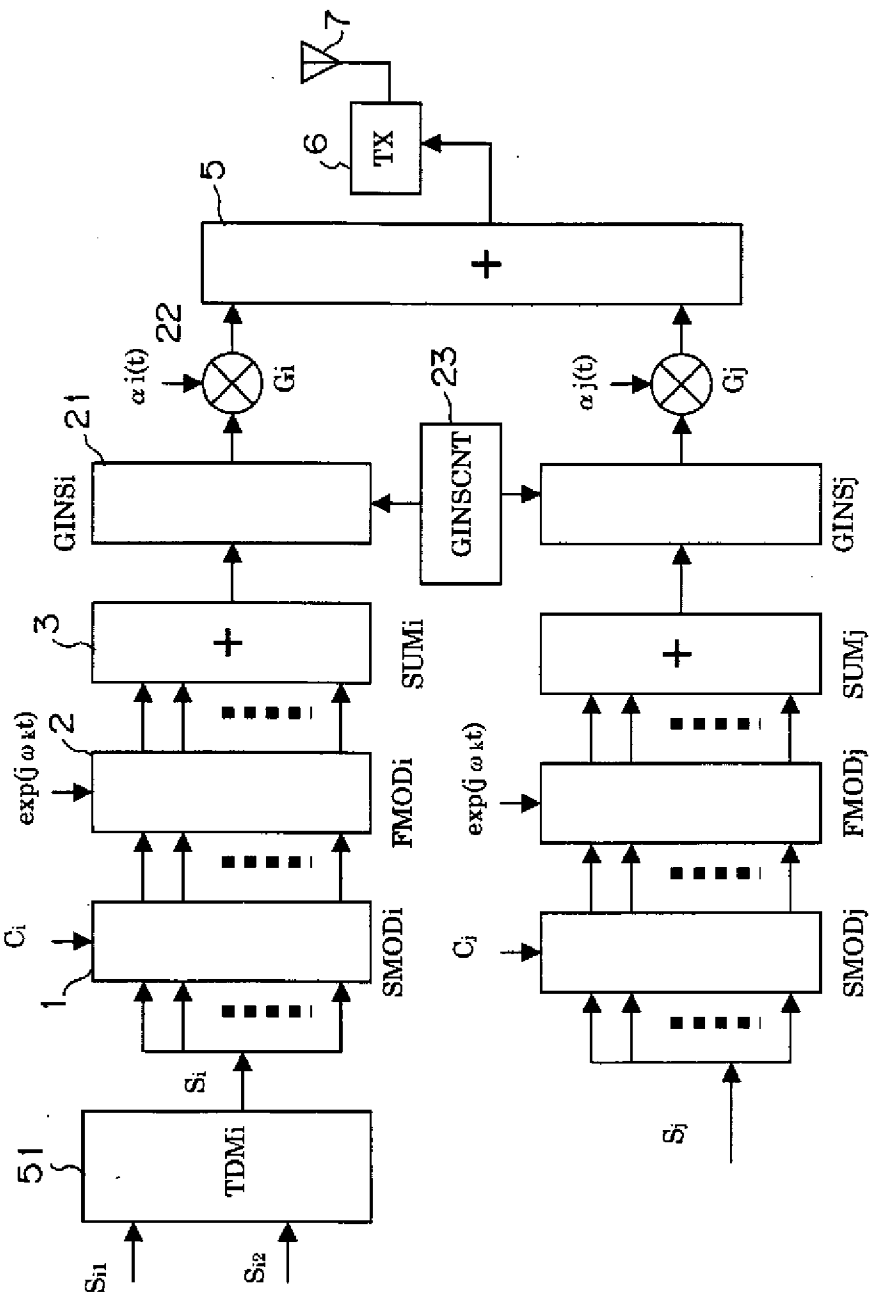

TX
GINSCNT
GINSi
GINSj
SUMi
SUMj
FMODi
FMODj
SMODi
SMODj
TDMi
Gi
Gj
α i(t)
α j(t)
exp(j ω kt)
Ci
Cj
Si
Sj
Si1
Si2
1
2
3
5
6
7
21
22
23
51

図15

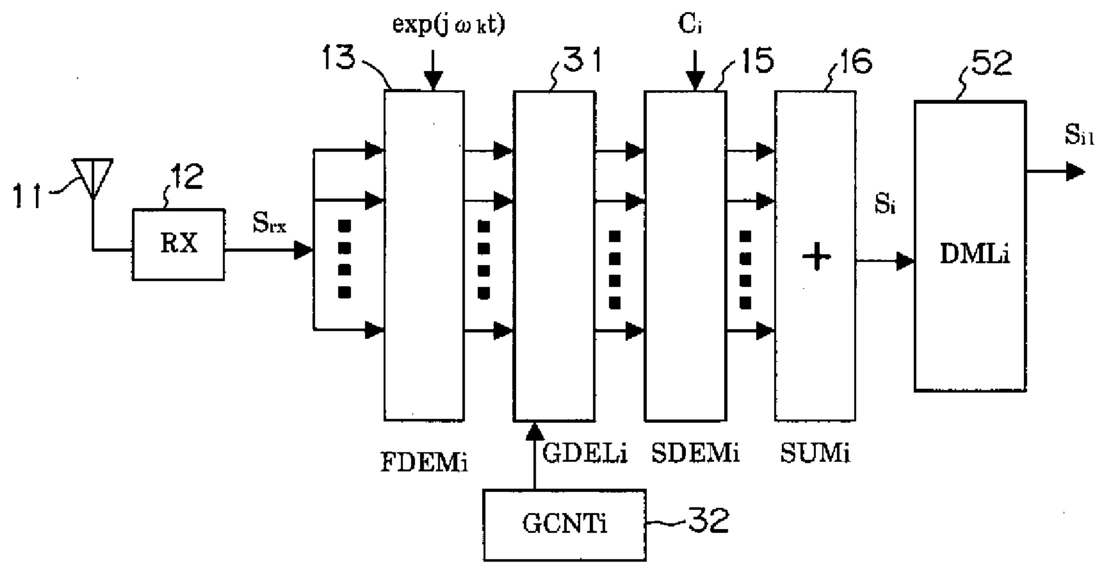

exp(jωkt)
Ci
13
31
15
16
52
11
12
RX
Srx
Si
Si1
+
DMLi
FDEMi
GDELi
SDEMi
SUMi
GCNTi
32

図16

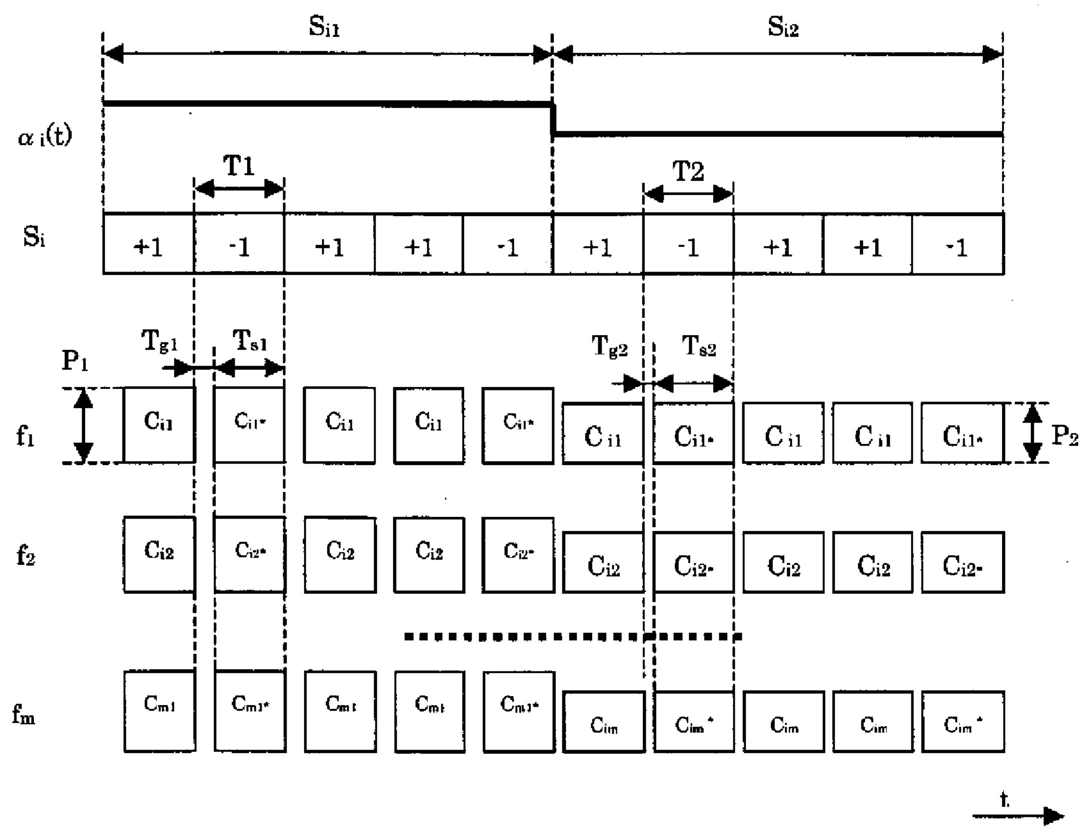

S_{i1}
S_{i2}
$\alpha_i(t)$
T1
T2
S_i
+1
-1
+1
+1
-1
+1
-1
+1
+1
-1
P_1
T_{g1}
T_{s1}
T_{g2}
T_{s2}
P_2
f_1
f_2
f_m
t

図17

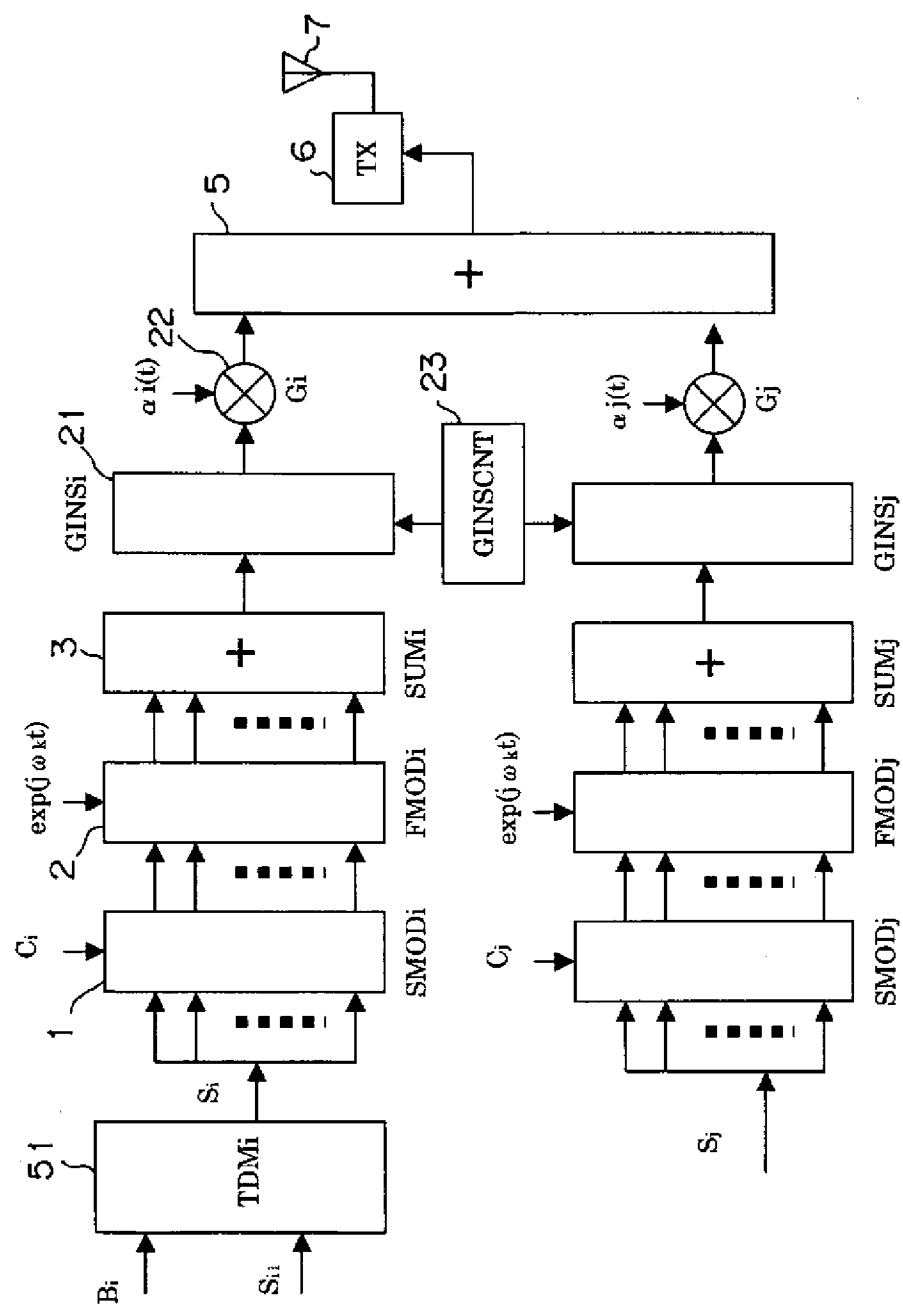

TDMi
SMODi
FMODi
SUMi
GINSi
GINSCNT
GINSj
SUMj
FMODj
SMODj
TX
Gi
Gj

図18

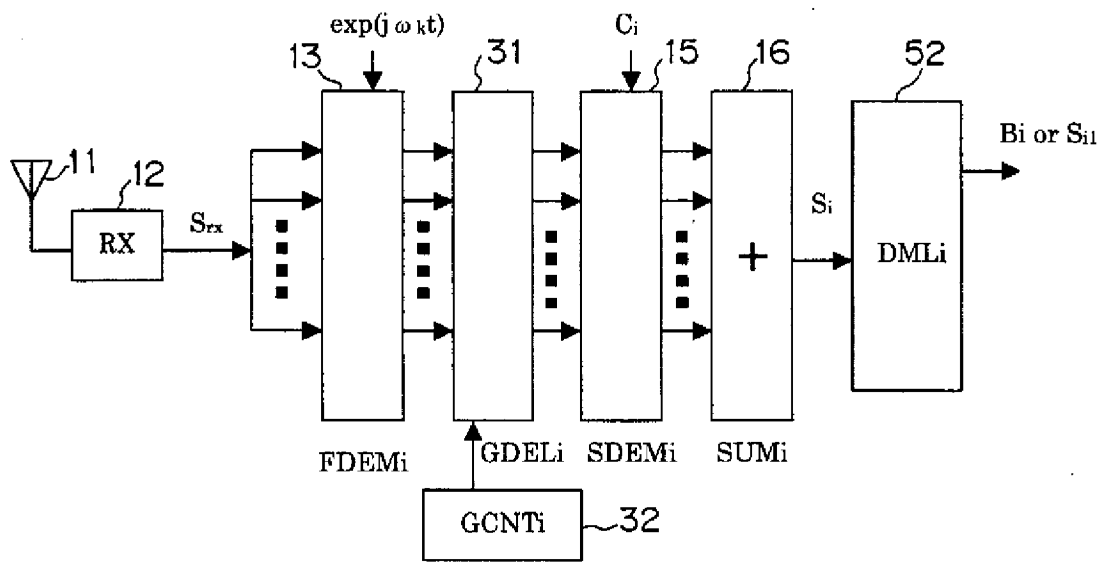

exp(j ω kt)
13
31
Ci
15
16
52
11
12
RX
Srx
+
Si
DMLi
Bi or Si1
FDEMi
GDELi
SDEMi
SUMi
GCNTi
32

図19

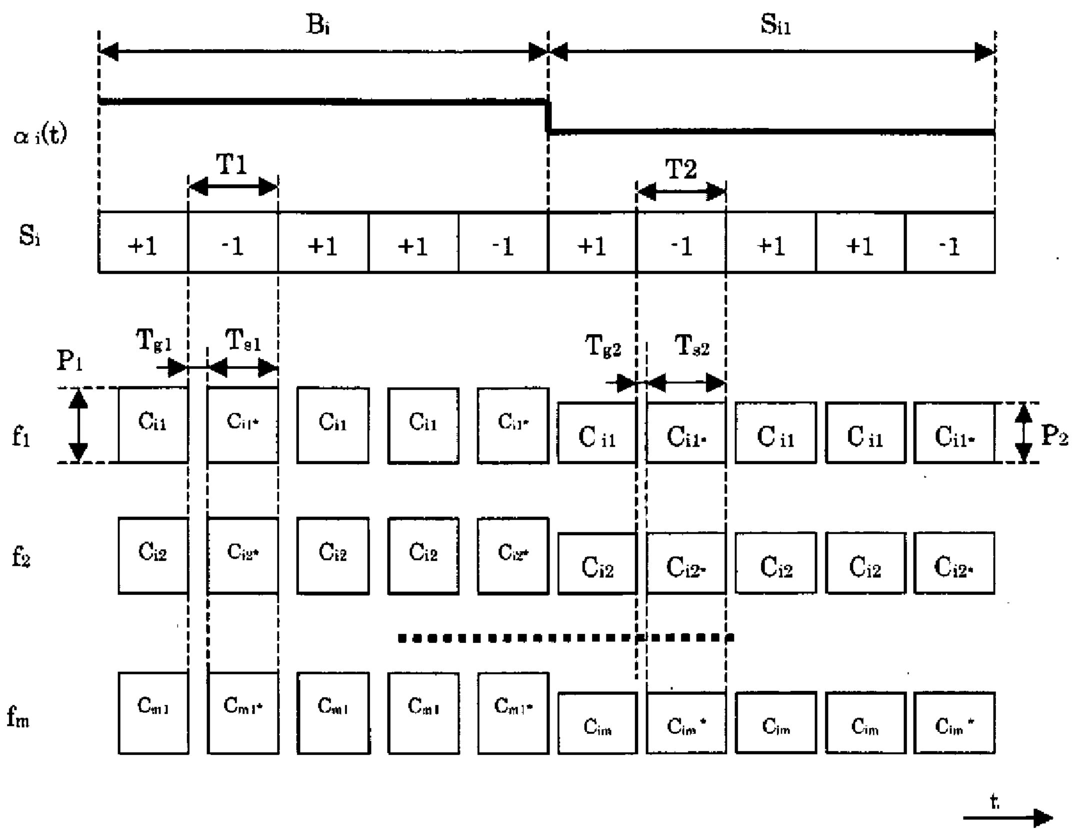


図20

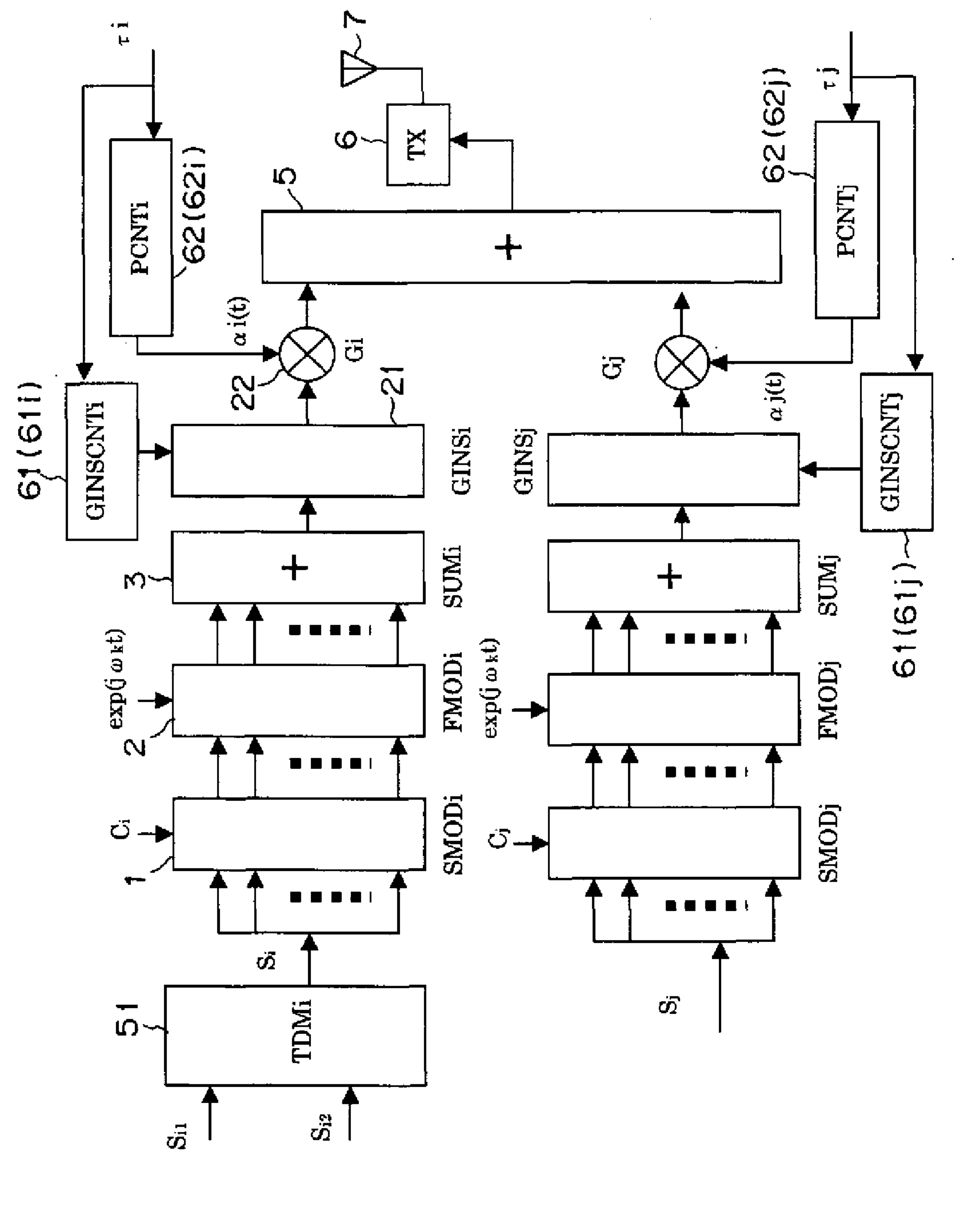
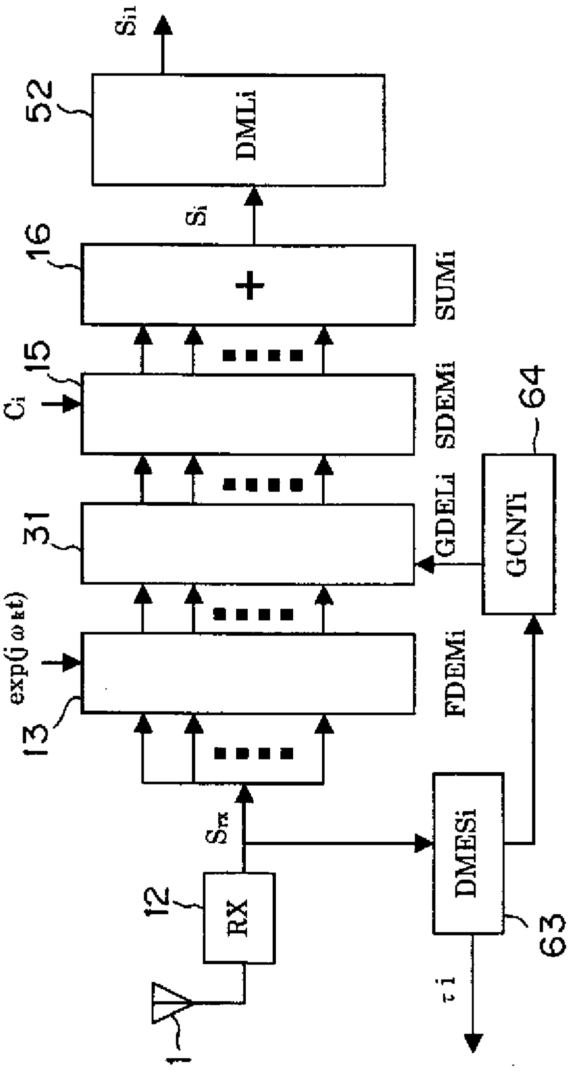

TDMi
SMODi
FMODi
SUMi
GINSi
GINSCNTi
PCNTi
TX
SMODj
FMODj
SUMj
GINSj
GINSCNTj
PCNTj
Gi
Gj
61(61i)
62(62i)
61(61j)
62(62j)

図21



Si1
52
DMLi
Si
16
+
SUMi
15
Ci
SDEMi
64
GCNTi
31
GDELi
exp(jωkt)
13
FDEMi
Srx
12
RX
11
DMESi
63
τi

図22

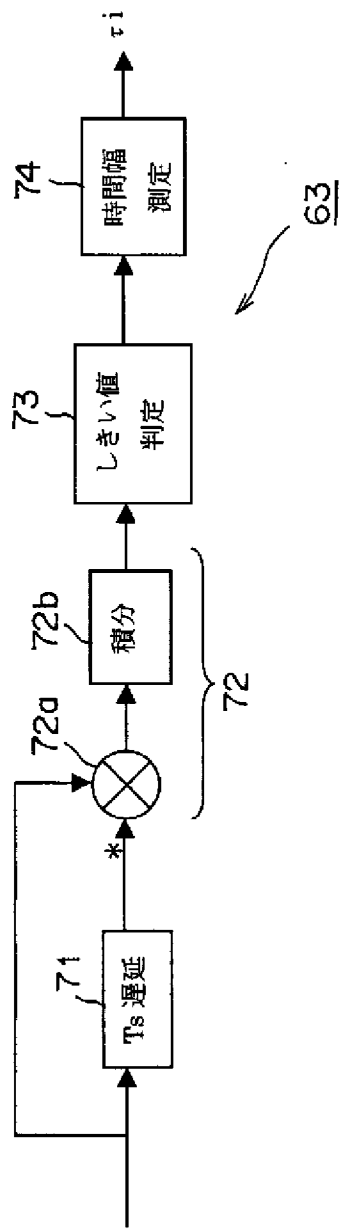


図23

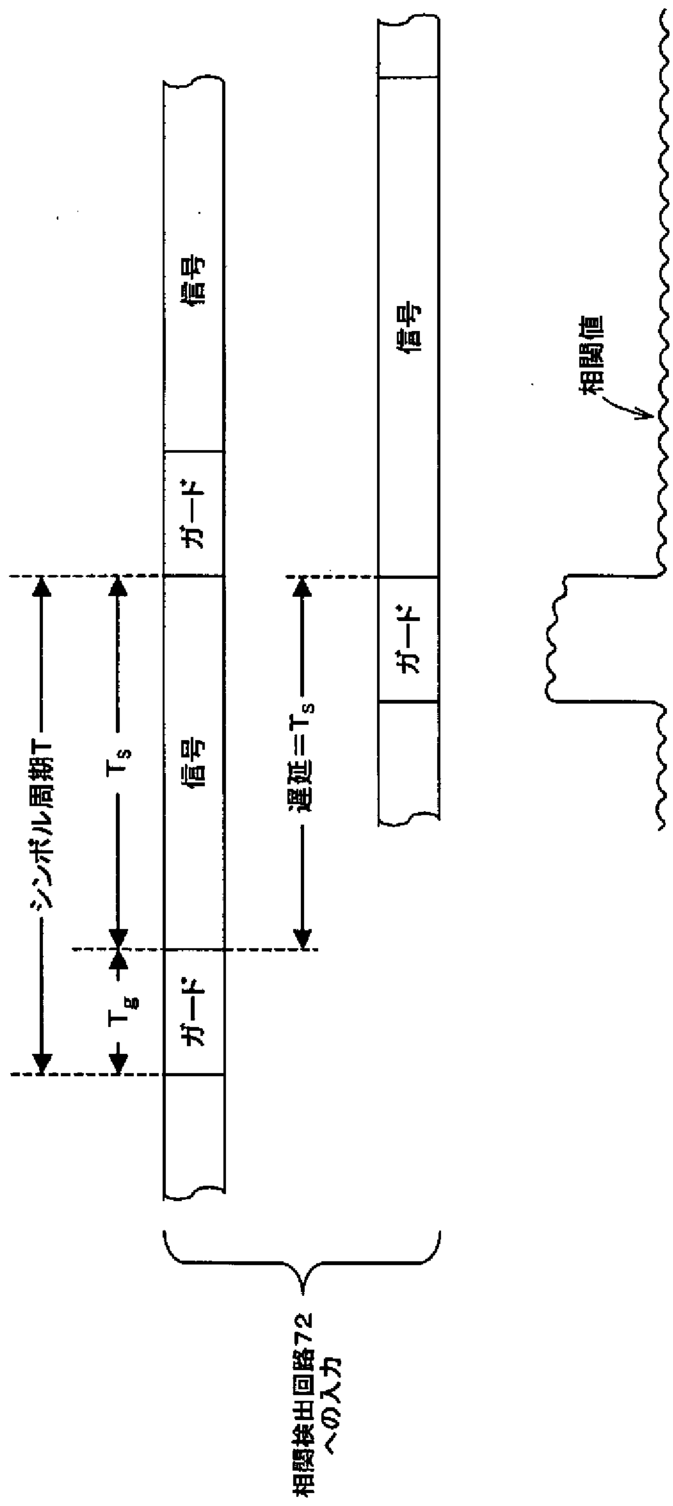


図24

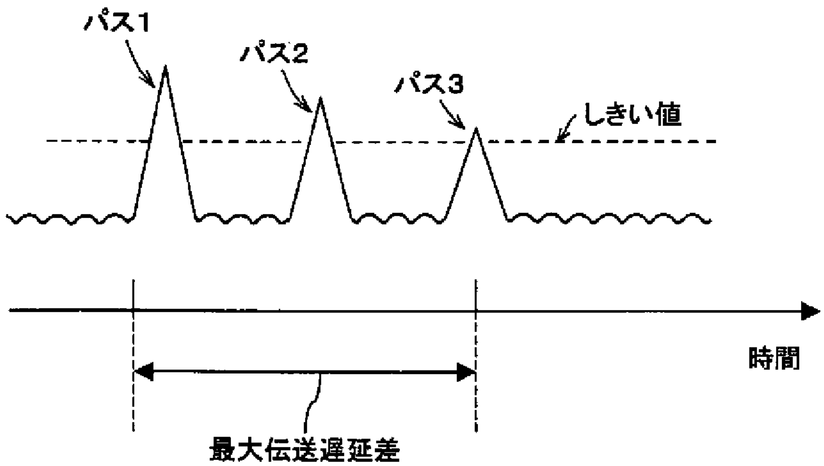


図25

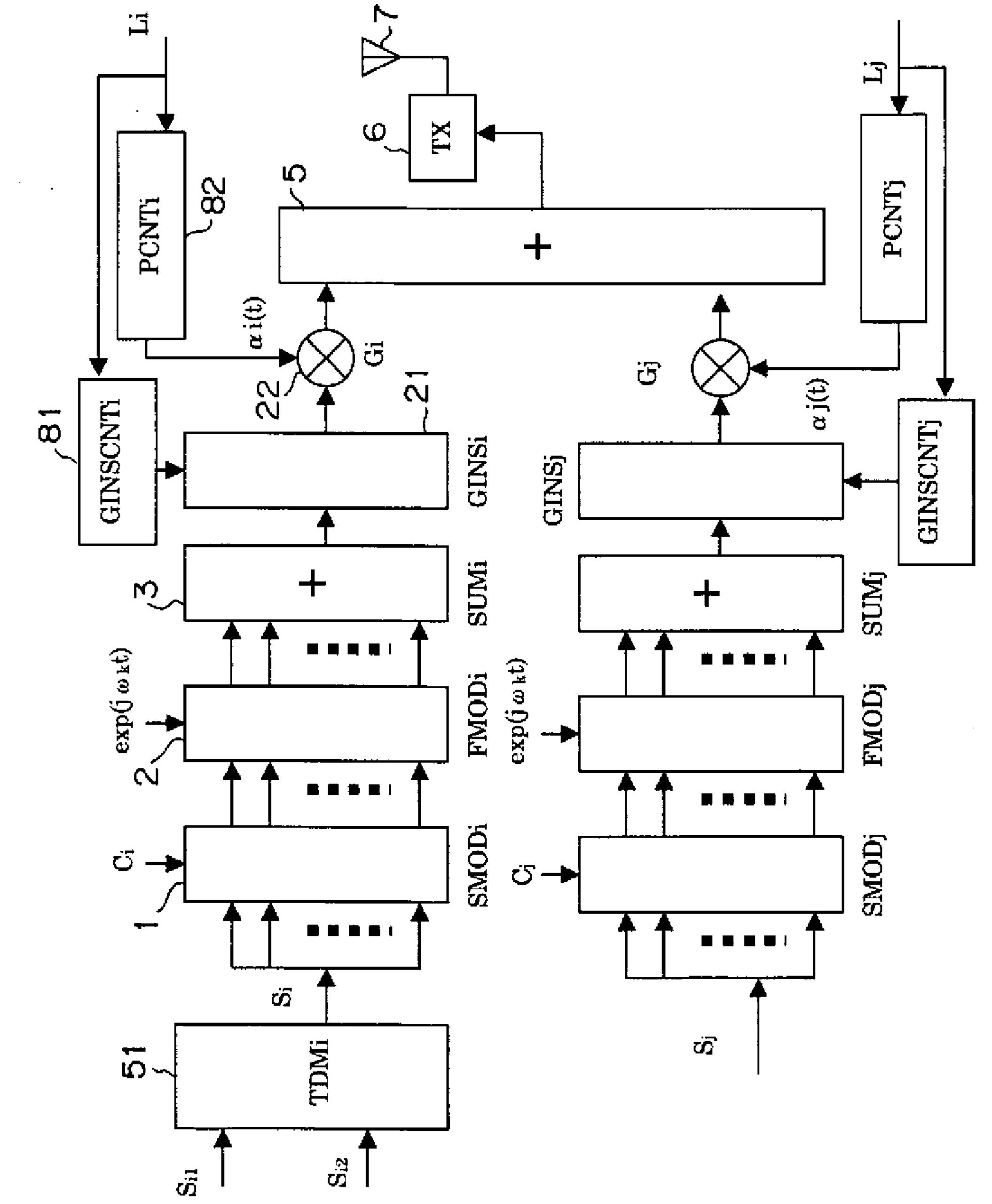

Li
Lj
7
6
TX
5
+
82
PCNTi
PCNTj
81
GINSCNTi
GINSCNTj
α i(t)
α j(t)
22
Gi
Gj
21
GINSi
GINSj
3
SUMi
SUMj
exp(jωkt)
2
FMODi
FMODj
Ci
Cj
1
SMODi
SMODj
Si
Sj
51
TDMi
Si1
Si2

図26

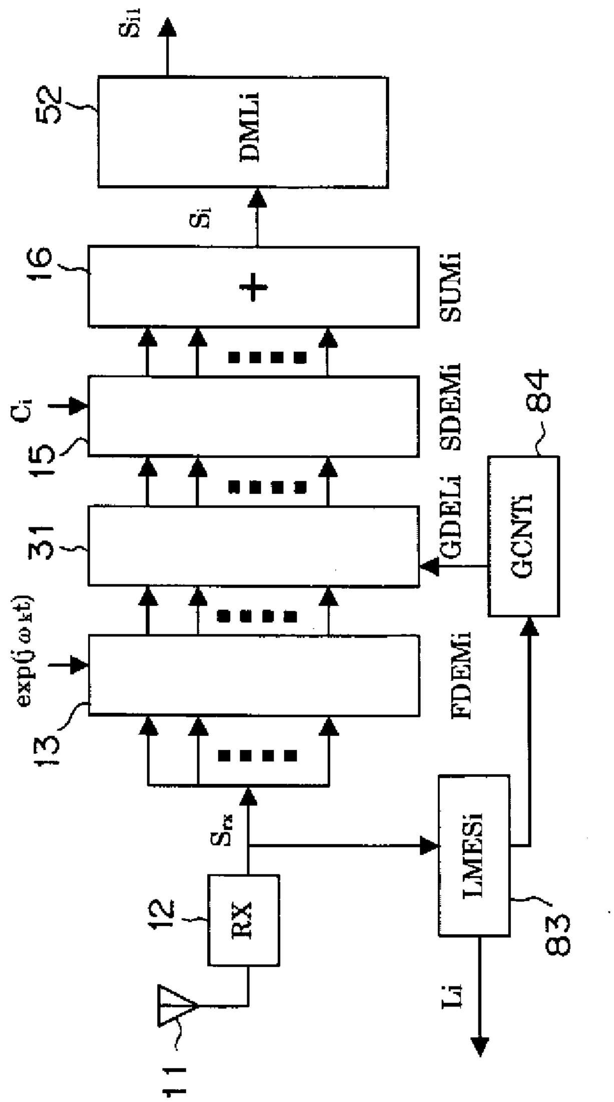

11
12
RX
S_{rx}
13
exp(jωkt)
FDEMi
31
GDELi
15
C_i
SDEMi
16
+
SUMi
S_i
52
DMLi
S_{i1}
LMESi
83
Li
GCNTi
84

図27

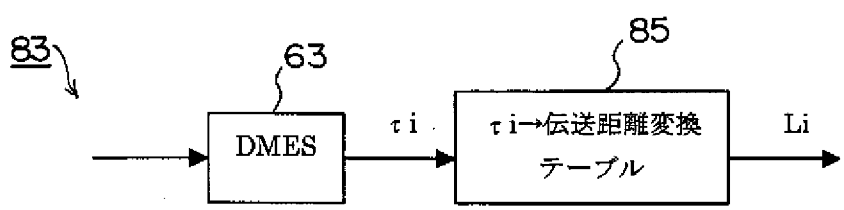


図28

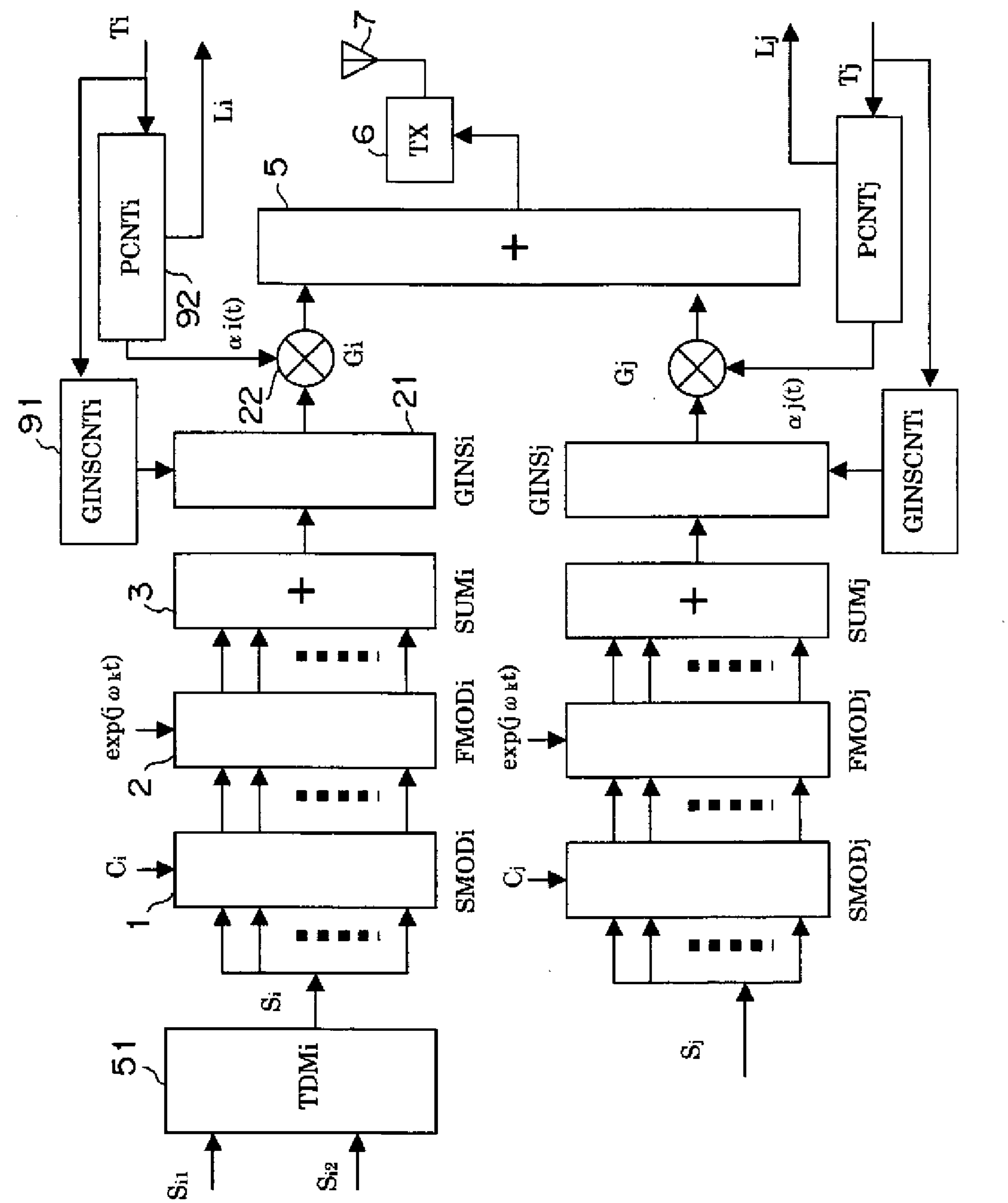

TDMi
51
S_{i1}
S_{i2}
S_i
SMODi
1
C_i
FMODi
2
$\exp(j\omega_k t)$
SUMi
3
+
GINSi
21
GINSCNTi
91
22
Gi
$\alpha i(t)$
PCNTi
92
Ti
Li
5
+
TX
6
7
S_j
SMODj
C_j
FMODj
$\exp(j\omega_k t)$
SUMj
+
GINSj
GINSCNTi
Gj
$\alpha j(t)$
PCNTj
Tj
Lj

図29

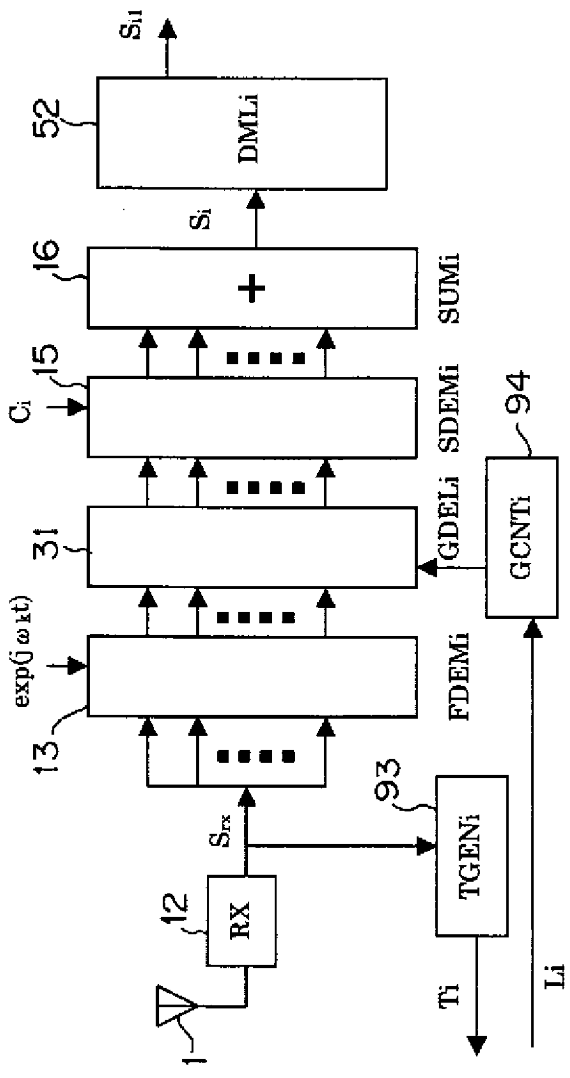

Si1
52
DMLi
Si
16
+
SUMi
15
Ci
SDEMi
94
GCNTi
31
GDELi
exp(jωkt)
FDEMi
13
93
TGENi
Srx
12
RX
11
Ti
Li

図30

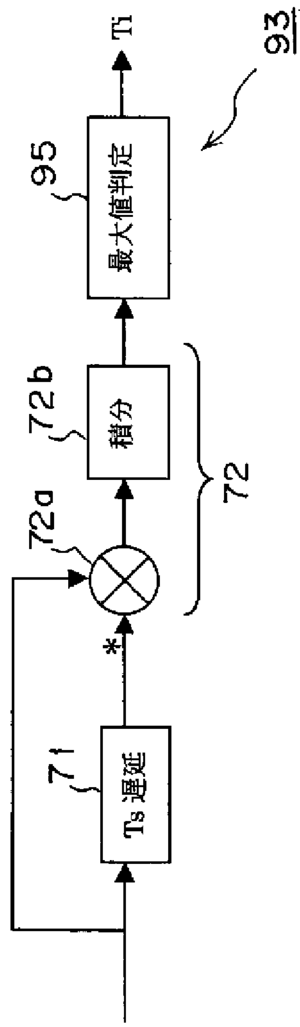
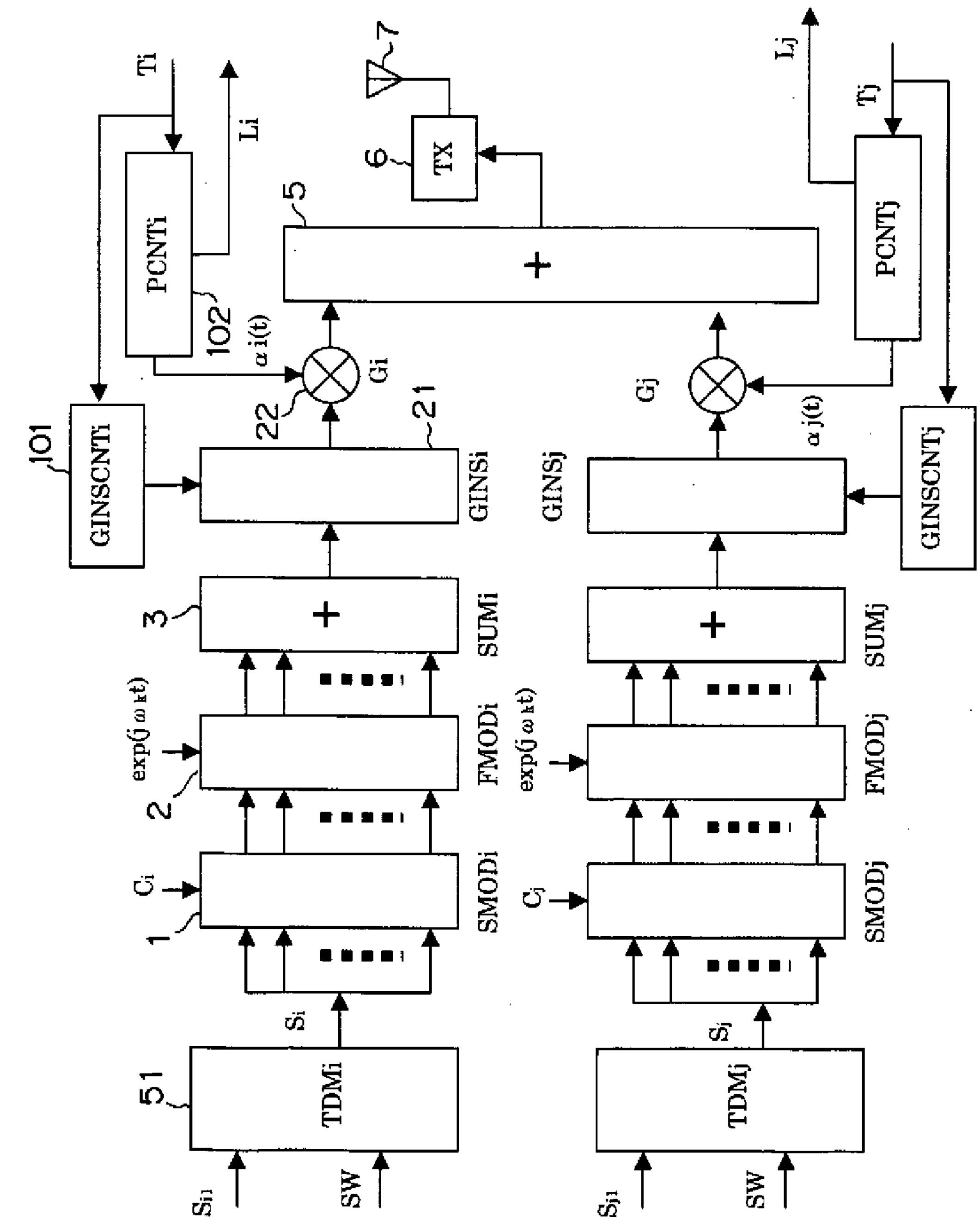


図31



Ti
Li
PCNTi
102
101
GINSCNTi
α i(t)
Gi
22
21
GINSi
3
SUMi
exp(j ωkt)
2
FMODi
Ci
1
SMODi
Si
51
TDMi
Si1
SW
7
6
TX
5
Lj
Tj
PCNTj
Gj
α j(t)
GINSj
GINSCNTj
SUMj
exp(j ωkt)
FMODj
Cj
SMODj
Sj
TDMj
Sj1
SW

図32

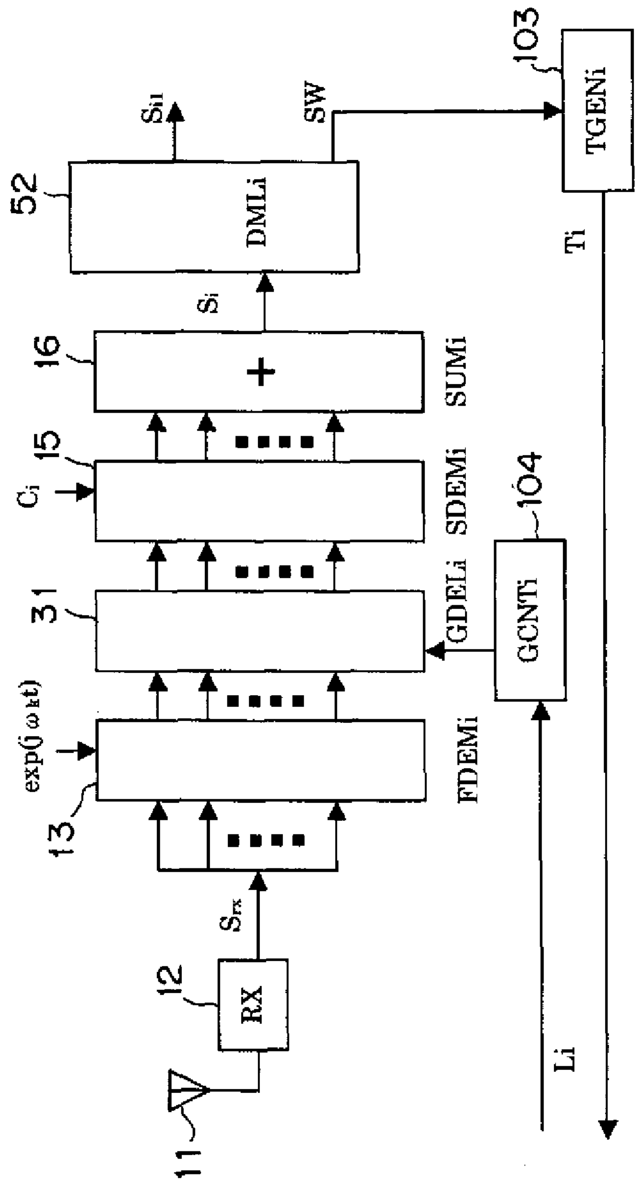

11
12
RX
S_{rx}
13
exp(jωkt)
FDEMi
31
GDELi
15
C_i
SDEMi
16
+
SUMi
S_i
52
DMLi
S_{i1}
SW
103
TGENi
Ti
104
GCNTi
Li

図33

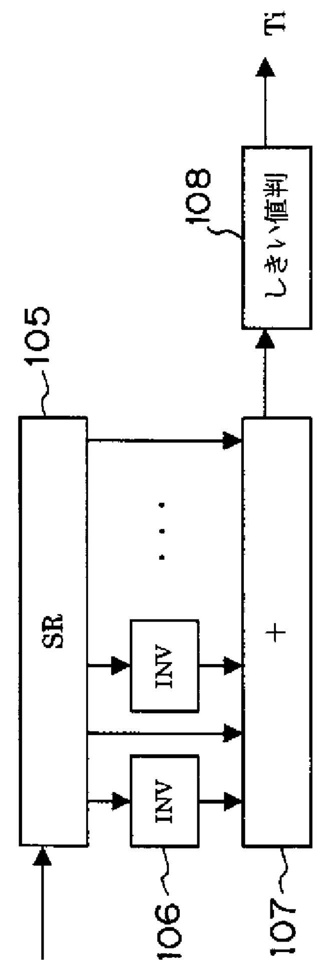


図34

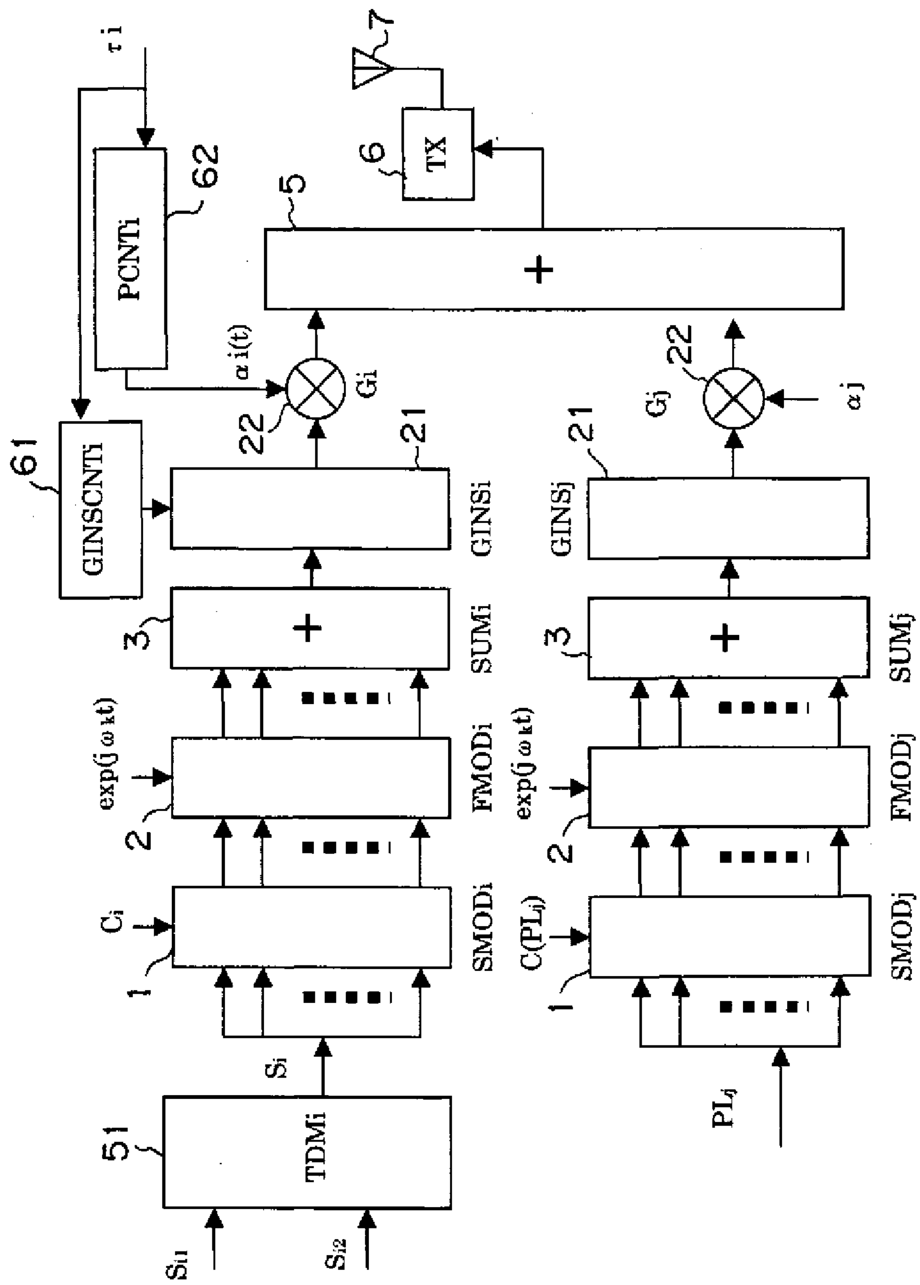

τi
PCNTi
62
GINSCNTi
61
7
TX
6
5
+
αi(t)
Gi
22
21
GINSi
SUMi
+
3
exp(jωkt)
FMODi
2
Ci
SMODi
1
Si
TDMi
51
Si1
Si2
22
αj
Gj
21
GINSj
SUMj
+
3
exp(jωkt)
FMODj
2
C(PLj)
SMODj
1
PLj

図35

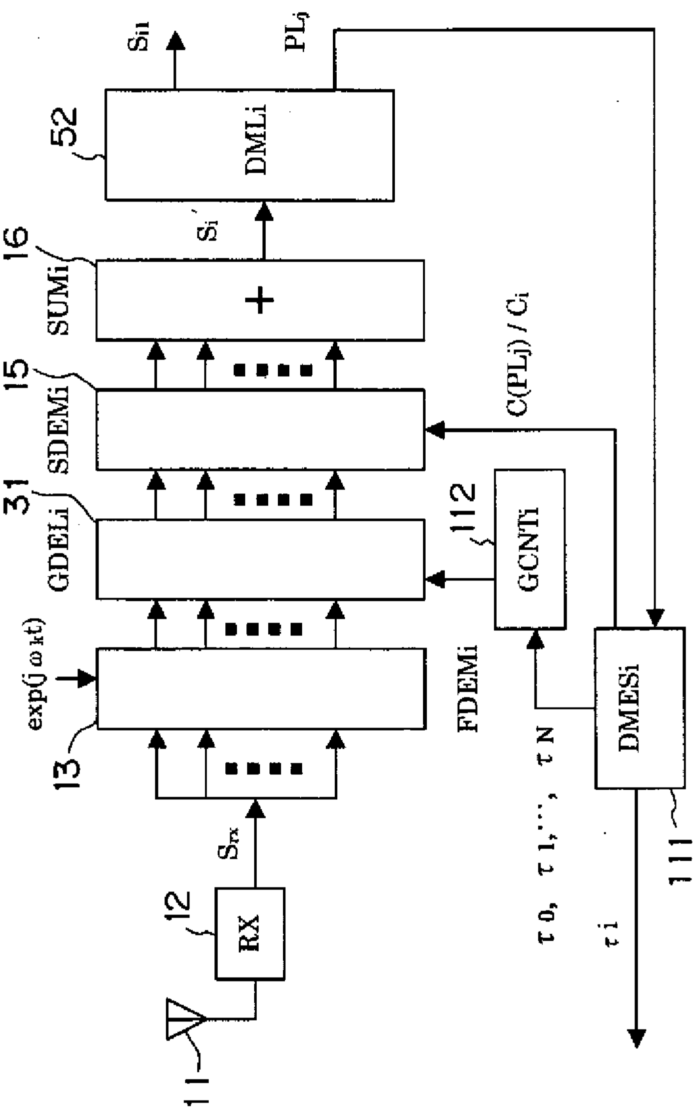

11
12
RX
S_{rx}
13
exp(jωkt)
FDEMi
31
GDELi
15
SDEMi
16
SUMi
+
S_i'
52
DMLi
S_{i1}
PL_j
$C(PL_j) / C_i$
112
GCNTi
DMESi
111
τ0, τ1,…, τN
τi

図36

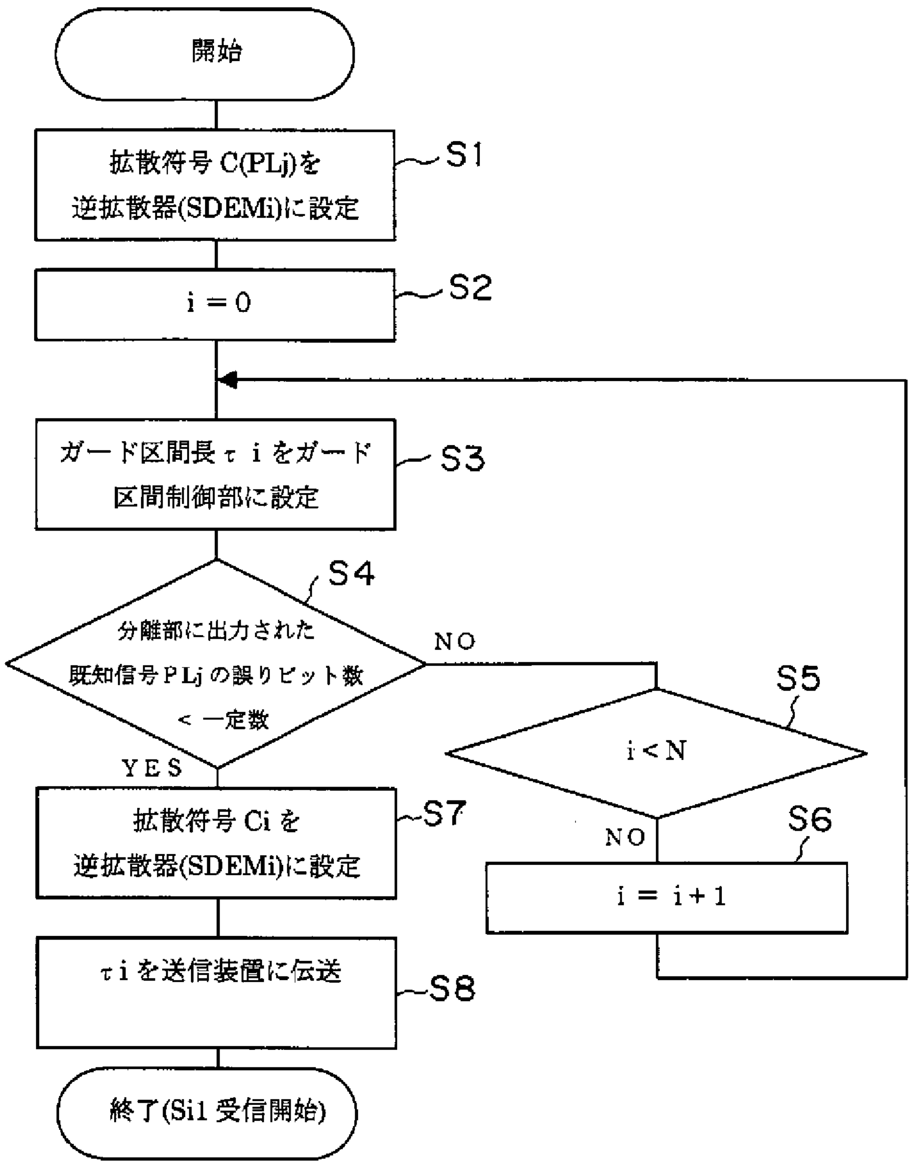

開始
拡散符号C(PLj)を
逆拡散器(SDEMi)に設定
S1
i＝0
S2
ガード区間長τｉをガード
区間制御部に設定
S3
S4
分離部に出力された
既知信号ＰＬｊの誤りビット数
＜一定数
NO
YES
S5
i＜N
NO
S6
i ＝ i＋1
拡散符号Ciを
逆拡散器(SDEMi)に設定
S7
τｉを送信装置に伝送
S8
終了(Si1受信開始)

図37

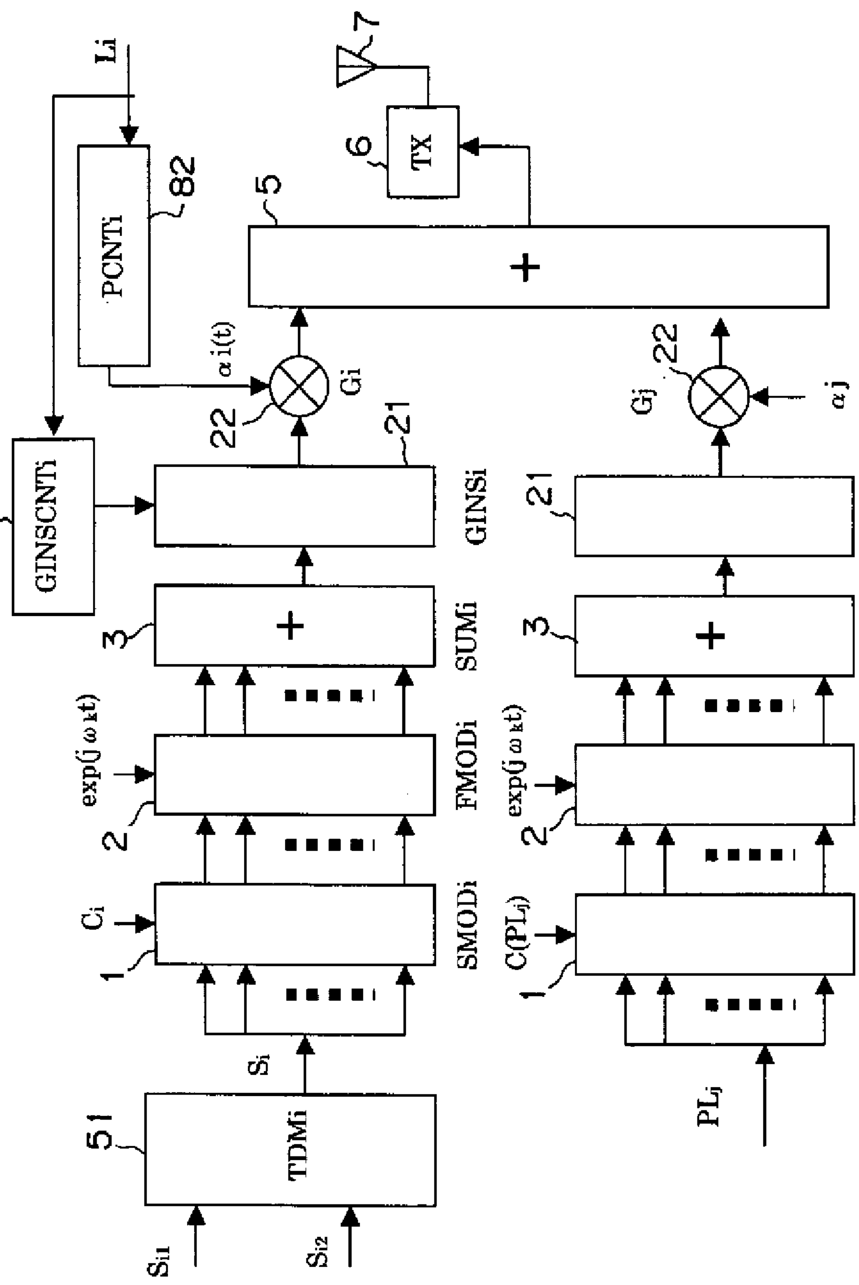

Li
PCNTi
82
7
TX
6
5
+
α i(t)
Gi
22
GINSCNTi
81
21
GINSi
+
SUMi
3
exp(j ω kt)
FMODi
2
Ci
SMODi
1
Si
TDMi
51
Si1
Si2
Gj
22
αj
21
GINSj
+
SUMj
3
exp(j ω kt)
FMODj
2
C(PLj)
SMODj
1
PLj

図38

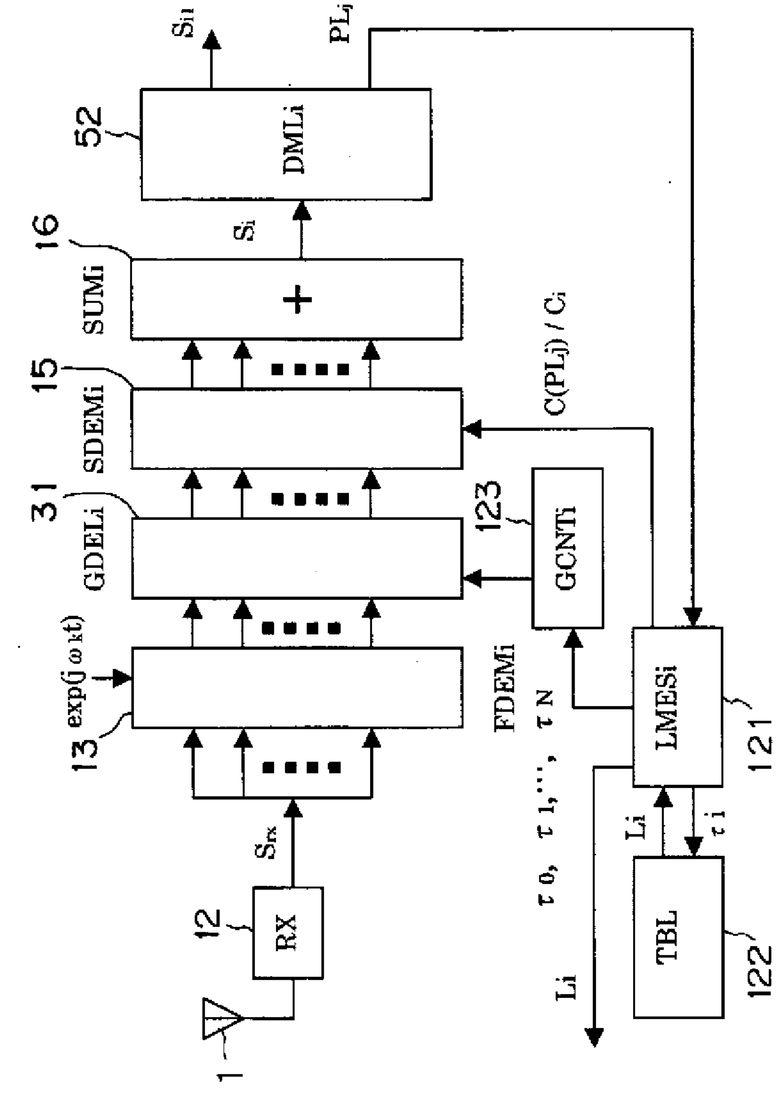

11
12
RX
S_{rx}
13
exp(jωkt)
FDEMi
31
GDELi
15
SDEMi
16
SUMi
+
S_i
52
DMLi
S_{i1}
PL_j
123
GCNTi
C(PLj) / C_i
121
LMESi
122
TBL
Li
τi
Li
τ0, τ1,…, τN

図39

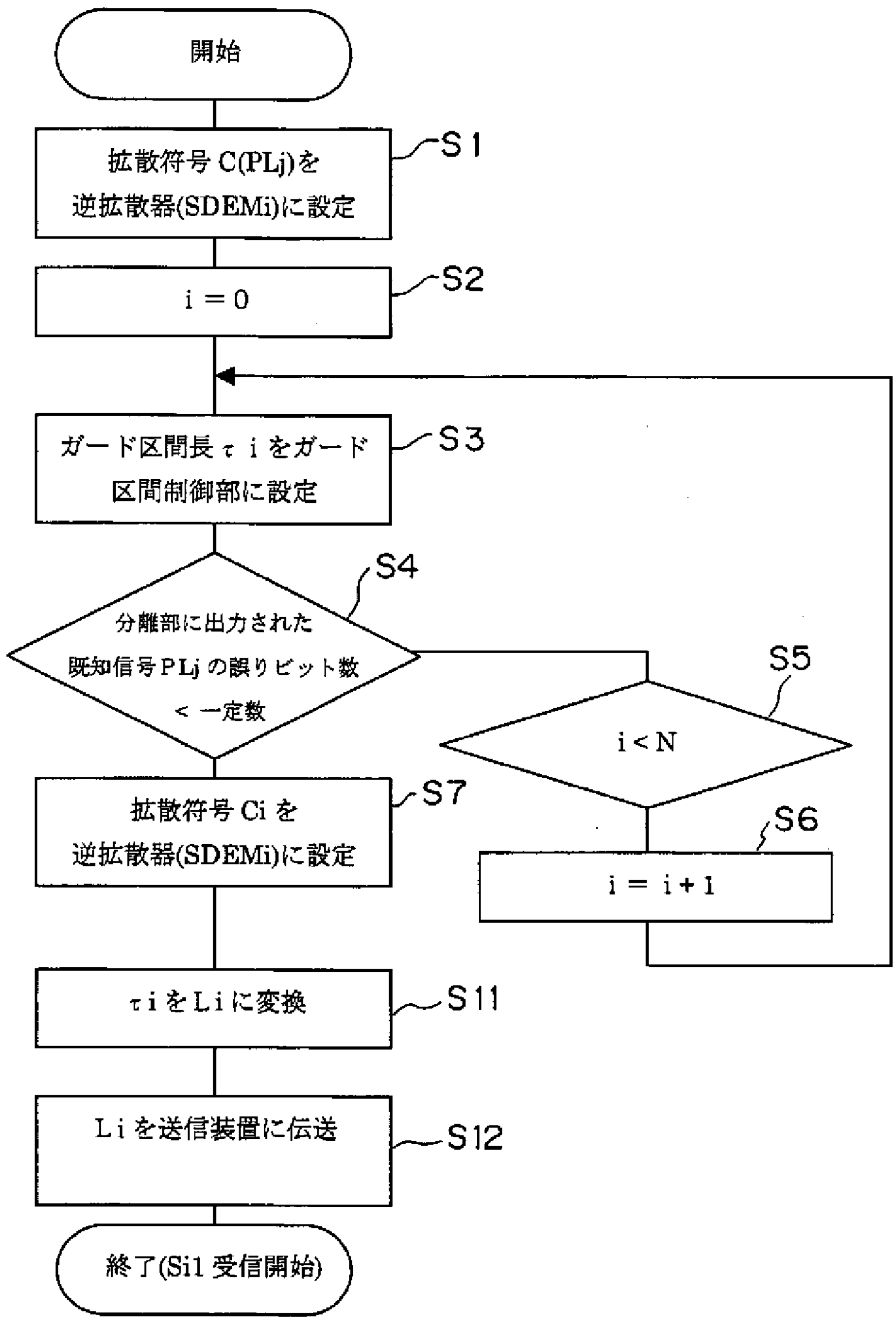

開始
拡散符号C(PLj)を
逆拡散器(SDEMi)に設定
S1
i＝0
S2
ガード区間長τiをガード
区間制御部に設定
S3
分離部に出力された
既知信号PLjの誤りビット数
＜一定数
S4
i＜N
S5
i＝i+1
S6
拡散符号Ciを
逆拡散器(SDEMi)に設定
S7
τiをLiに変換
S11
Liを送信装置に伝送
S12
終了(Si1受信開始)

図40

INTERNATIONAL SEARCH REPORT

International application No.
PCT/JP01/10357

A. CLASSIFICATION OF SUBJECT MATTER
Int.Cl[7] H04J11/00

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
Int.Cl[7] H04J11/00

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched
Jitsuyo Shinan Koho 1926-2000
Kokai Jitsuyo Shinan Koho 1971-2000

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	EP 1035693 A (Matsushita Electric Industrial Co., Ltd.), 13 September 2000 (13.09.2000), page 43, lines 7 to 40; Fig.146 & JP 7-99522 A, page 36, left column, line 35 to right column, line 32	1,3-5,11,13, 16,18-20
A		7-9
X	JP 2000-165342 A (Matsushita Electric Ind. Co., Ltd.), 16 June 2000 (16.06.2000), page 3, left column, line 42 to right column, line 17	1,2,6,10-12, 14-17
A	(Family: none)	7-9
X	EP 1014639 A (Matsushita Electric Industrial Co., Ltd.), 28 June 2000 (28.06.2000), page 4, lines 13 to 18 & JP 2000-244441 A, page 6, right column, lines 22 to 34 & CN 1260649 A & KR 2000052538 A	1,6,10,11, 14-16
A		7-9
A	JP 2001-111519 A (Matsushita Electric Ind. Co., Ltd.), 20 February 2001 (20.04.2001), page 3, right column, lines 9 to 17	1-20

☒ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:
"A" document defining the general state of the art which is not considered to be of particular relevance
"E" earlier document but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed
"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
29 January, 2002 (29.01.02)	05 February, 2002 (05.02.02)
Name and mailing address of the ISA/ Japanese Patent Office	Authorized officer
Facsimile No.	Telephone No.

INTERNATIONAL SEARCH REPORT

International application No.
PCT/JP01/10357

C (Continuation). DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
	(Family: none)	
A	JP 11-196062 A (Advanced Digital Television Broadcasting), 21 July 1999 (21.07.1991), page 2, right column, lines 37 to 42 (Family: none)	1-20

国際調査報告

国際出願番号　ＰＣＴ／ＪＰ０１／１０３５７

Ａ．　発明の属する分野の分類（国際特許分類（ＩＰＣ））
Ｉｎｔ．Ｃｌ⁷　Ｈ０４Ｊ１１／００

Ｂ．　調査を行った分野
調査を行った最小限資料（国際特許分類（ＩＰＣ））
Ｉｎｔ．Ｃｌ⁷　Ｈ０４Ｊ１１／００

最小限資料以外の資料で調査を行った分野に含まれるもの
日本国実用新案公報　　１９２６－２０００
日本国公開実用新案公報　　１９７１－２０００

国際調査で使用した電子データベース（データベースの名称、調査に使用した用語）

Ｃ．　関連すると認められる文献

引用文献のカテゴリー＊	引用文献名　及び一部の箇所が関連するときは、その関連する箇所の表示	関連する請求の範囲の番号
X A	ＥＰ　１０３５６９３　Ａ（Ｍａｔｓｕｓｈｉｔａ　Ｅｌｅｃｔｒｉｃ　Ｉｎｄｕｓｔｒｉａｌ　Ｃｏ．，　Ｌｔｄ．），２０００．０９．１３，第４３頁第７行目－第４０行目，ＦＩＧ．１４６＆　ＪＰ　７－９９５２２　Ａ，第３６頁左欄第３５行目－右欄第３２行目	1, 3-5, 11, 13, 16, 18-20 7-9

☒　Ｃ欄の続きにも文献が列挙されている。　　☐　パテントファミリーに関する別紙を参照。

＊　引用文献のカテゴリー
「Ａ」特に関連のある文献ではなく、一般的技術水準を示すもの
「Ｅ」国際出願日前の出願または特許であるが、国際出願日以後に公表されたもの
「Ｌ」優先権主張に疑義を提起する文献又は他の文献の発行日若しくは他の特別な理由を確立するために引用する文献（理由を付す）
「Ｏ」口頭による開示、使用、展示等に言及する文献
「Ｐ」国際出願日前で、かつ優先権の主張の基礎となる出願の日の後に公表された文献
「Ｔ」国際出願日又は優先日後に公表された文献であって出願と矛盾するものではなく、発明の原理又は理論の理解のために引用するもの
「Ｘ」特に関連のある文献であって、当該文献のみで発明の新規性又は進歩性がないと考えられるもの
「Ｙ」特に関連のある文献であって、当該文献と他の１以上の文献との、当業者にとって自明である組合せによって進歩性がないと考えられるもの
「＆」同一パテントファミリー文献

国際調査を完了した日　２９．０１．０２	国際調査報告の発送日　05.02.02
国際調査機関の名称及びあて先 日本国特許庁（ＩＳＡ／ＪＰ） 郵便番号１００－８９１５ 東京都千代田区霞が関三丁目４番３号	特許庁審査官（権限のある職員）　５Ｋ　９６４７ 高野　洋 電話番号　０３－３５８１－１１０１　内線　３５５５

Ｃ（続き）．　関連すると認められる文献

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A		7-9
X	ＥＰ　１０１４６３９　Ａ（Ｍａｔｓｕｓｈｉｔａ　Ｅｌｅｃｔｒｉｃ　Ｉｎｄｕｓｔｒｉａｌ　Ｃｏ．，　Ｌｔｄ．），２０００．０６．２８，第４頁第１３行目－第１８行目 ＆　ＪＰ　２０００－２４４４４１　Ａ，第６頁右欄第２２行目－第３４行目 ＆　ＣＮ　１２６０６４９　Ａ ＆　ＫＲ　２０００００５２５３８　Ａ	1, 6, 10, 11, 14-16
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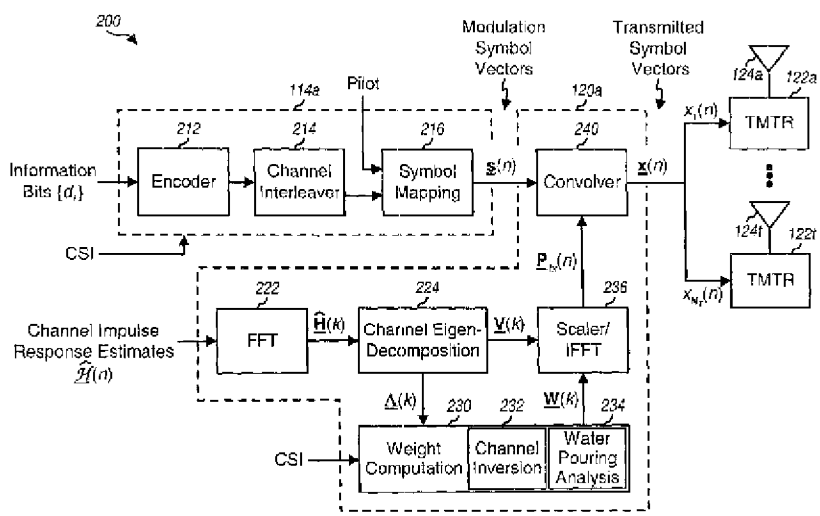
(54) **Title:** SIGNAL PROCESSING WITH CHANNEL EIGENMODE DECOMPOSITION AND CHANNEL INVERSION FOR MIMO SYSTEMS



(57) **Abstract:** Techniques for processing a data transmission at a transmitter and receiver, which use channel eigen-decomposition, channel inversion, and (optionally) "water-pouring". At the transmitter, (1) channel eigen-decomposition is performed to determine eigenmodes of a MIMO channel and to derive a first set of steering vectors, (2) channel inversion is performed to derive weights (*e.g.*, one set for each eigenmode) used to minimize ISI distortion, and (3) water-pouring may be performed to derive scaling values indicative of the transmit powers allocated to the eigenmodes. The first set of steering vectors, weights, and scaling values are used to derive a pulse-shaping matrix, which is used to precondition modulation symbols prior to transmission. At the receiver, channel eigen-decomposition is performed to derive a second set of steering vectors, which are used to derive a pulse-shaping matrix used to condition received symbols such that orthogonal symbol streams are recovered.

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SIGNAL PROCESSING WITH CHANNEL EIGENMODE DECOMPOSITION AND CHANNEL INVERSION FOR MIMO SYSTEMS

BACKGROUND

Field

[1001] The present invention relates generally to data communication, and more specifically to techniques for performing signal processing with channel eigenmode decomposition and channel inversion for multiple-input multiple-output (MIMO) communication systems.

Background

[1002] A multiple-input multiple-output (MIMO) communication system employs multiple (N_T) transmit antennas and multiple (N_R) receive antennas for data transmission. A MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_S independent channels, with $N_S \leq \min\{N_T, N_R\}$. Each of the N_S independent channels is also referred to as a spatial subchannel of the MIMO channel and corresponds to a dimension. The MIMO system can provide improved performance (e.g., increased transmission capacity) if the additional dimensionalities created by the multiple transmit and receive antennas are utilized.

[1003] The spatial subchannels of a wideband MIMO system may encounter different channel conditions due to various factors such as fading and multipath. Each spatial subchannel may thus experience frequency selective fading, which is characterized by different channel gains at different frequencies (i.e., different frequency bins or subbands) of the overall system bandwidth. With frequency selective fading, each spatial subchannel may achieve different signal-to-noise-and-interference ratios (SNRs) for different frequency bins. Consequently, the number of information bits per modulation symbol (or data rate) that may be transmitted at different frequency bins of each spatial subchannel for a particular level of performance (e.g., 1% packet error rate) may be different from bin to bin. Moreover, because the channel conditions

typically vary with time, the supported data rates for the bins of the spatial subchannels also vary with time.

[1004] To combat frequency selective fading in a wideband channel, orthogonal frequency division multiplexing (OFDM) may be used to effectively partition the system bandwidth into a number of (N_F) subbands (which may also be referred to as frequency bins or subchannels). With OFDM, each frequency subchannel is associated with a respective subcarrier upon which data may be modulated. For a MIMO system that utilizes OFDM (i.e., a MIMO-OFDM system), each frequency subchannel of each spatial subchannel may be viewed as an independent transmission channel.

[1005] A key challenge in a coded communication system is the selection of the appropriate data rates and coding and modulation schemes to be used for a data transmission based on channel conditions. The goal of this selection process is to maximize throughput while meeting quality objectives, which may be quantified by a particular packet error rate (PER), certain latency criteria, and so on.

[1006] One straightforward technique for selecting data rates and coding and modulation schemes is to "bit load" each transmission channel in the MIMO-OFDM system according to its transmission capability, which may be quantified by the channel's short-term average SNR. However, this technique has several major drawbacks. First, coding and modulating individually for each transmission channel can significantly increase the complexity of the processing at both the transmitter and receiver. Second, coding individually for each transmission channel may greatly increase coding and decoding delay. And third, a high feedback rate would be needed to send channel state information (CSI) indicative of the channel conditions (e.g., the gain, phase, and SNR) of each transmission channel.

[1007] For a MIMO system, transmit power is another parameter that may be manipulated to maximize throughput. In general, the overall throughput of the MIMO system may be increased my allocating more transmit power to transmission channels with greater transmission capabilities. However, allocating different amounts of transmit power to different frequency bins of a given spatial subchannel tends to exaggerate the frequency selective nature of the spatial subchannel. It is well known that frequency selective fading causes inter-symbol interference (ISI), which is a phenomenon whereby each symbol in a received signal acts as distortion to subsequent symbols in the received signal. The ISI distortion degrades performance by impacting

the ability to correctly detect the received symbols. To mitigate the deleterious effects of ISI, equalization of the received symbols would need to be performed at the receiver. Thus, a major drawback in frequency-domain power allocation is the additional complexity at the receiver to combat the resultant additional ISI distortion.

[1008] There is therefore a need in the art for techniques to achieve high overall throughput in a MIMO system without having to individually code each transmission channel and which mitigate the deleterious effects of ISI.

SUMMARY

[1009] Techniques are provided herein for processing a data transmission at a transmitter and a receiver of a MIMO system such that high performance (e.g., high overall throughput) is achieved. In an aspect, a time-domain implementation is provided which uses frequency-domain channel eigen-decomposition, channel inversion, and (optionally) "water-pouring" results to derive pulse-shaping and beam-steering solutions for the transmitter and receiver.

[1010] Channel eigen-decomposition is performed at the transmitter to determine the eigenmodes (i.e., the spatial subchannels) of a MIMO channel and to obtain a first set of steering vectors, which are used to precondition modulation symbols prior to transmission over the MIMO channel. Channel eigen-decomposition may be performed based on an estimated channel response matrix, which is an estimate of the (time-domain or frequency-domain) channel response of the MIMO channel. Channel eigen-decomposition is also performed at the receiver to obtain a second set of steering vectors, which are used to condition received symbols such that orthogonal symbol streams are recovered at the receiver.

[1011] Channel inversion is performed at the transmitter to derive weights, which are used to minimize or reduce the amount of ISI distortion at the receiver. In particular, the channel inversion may be performed for each eigenmode used for data transmission. One set of weights may be derived for each eigenmode based on the estimated channel response matrix for the MIMO channel and is used to invert the frequency response of the eigenmode.

[1012] Water-pouring analysis may (optionally) be used to more optimally allocate the total available transmit power to the eigenmodes of the MIMO channel. In particular, eigenmodes with greater transmission capabilities may be allocated more

transmit power, and eigenmodes with transmission capabilities below a particular threshold may be omitted from use (e.g., by allocating these bad eigenmodes with zero transmit power). The transmit power allocated to each eigenmode then determines the data rate and possibly the coding and modulation scheme to be used for the eigenmode.

[1013] At the transmitter, data is initially processed (e.g., coded and modulated) in accordance with a particular processing scheme to provide a number of modulation symbol streams (e.g., one modulation symbol stream for each eigenmode). An estimated channel response matrix for the MIMO channel is obtained (e.g., from the receiver) and decomposed (e.g., in the frequency domain, using channel eigen-decomposition) to obtain a set of matrices of right eigen-vectors and a set of matrices of singular values. A number of sets of weights are then derived based on the matrices of singular values, with each set of weights being used to invert the frequency response of a respective eigenmode used for data transmission. Water-pouring analysis may also be performed based on the matrices of singular values to obtain a set of scaling values indicative of the transmit powers allocated to the eigenmodes. A pulse-shaping matrix for the transmitter is then derived based on the matrices of right eigen-vectors, the weights, and the scaling values (if available). The pulse-shaping matrix comprises steering vectors, which are used to precondition the streams of modulation symbols to obtain streams of preconditioned symbols to be transmitted over the MIMO channel.

[1014] At the receiver, the estimated channel response matrix is also obtained (e.g., based on pilot symbols sent from the transmitter) and decomposed to obtain a set of matrices of left eigen-vectors. A pulse-shaping matrix for the receiver is then derived based on the matrices of left eigen-vectors and used to condition a number of received symbol streams to obtain a number of recovered symbol streams. The recovered symbols are further processed (e.g., demodulated and decoded) to recover the transmitted data.

[1015] Various aspects and embodiments of the invention are described in further detail below. The invention further provides methods, digital signal processors, transmitter and receiver units, and other apparatuses and elements that implement various aspects, embodiments, and features of the invention, as described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

[1016] The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

[1017] FIG. 1 is a block diagram of an embodiment of a transmitter system and a receiver system in a MIMO system;

[1018] FIG. 2 is a block diagram of a transmitter unit within the transmitter system;

[1019] FIGS. 3A and 3B are diagrams that graphically illustrate the derivation of the weights used to invert the frequency response of each eigenmode of a MIMO channel;

[1020] FIG. 4 is a flow diagram of a process for allocating the total available transmit power to the eigenmodes of the MIMO channel;

[1021] FIGS. 5A and 5B are diagrams that graphically illustrate the allocation of the total transmit power to three eigenmodes in an example MIMO system;

[1022] FIG. 6 is a flow diagram of an embodiment of the signal processing at the transmitter unit;

[1023] FIG. 7 is a block diagram of a receiver unit within the receiver system; and

[1024] FIG. 8 is a flow diagram of an embodiment of the signal processing at the receiver unit.

DETAILED DESCRIPTION

[1025] The techniques described herein for processing a data transmission at a transmitter and receiver may be used for various wireless communication systems. For clarity, various aspects and embodiments of the invention are described specifically for a multiple-input multiple-output (MIMO) communication system.

[1026] A MIMO system employs multiple (N_T) transmit antennas and multiple (N_R) receive antennas for data transmission. A MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_S independent channels, with $N_S \leq \min\{N_T, N_R\}$. Each of the N_S independent channels is also referred to as a spatial subchannel of the MIMO channel. The number of spatial subchannels is determined by the number of eigenmodes for the MIMO channel, which in turn is dependent on a

channel response matrix that describes the response between the N_T transmit and N_R receive antennas.

[1027] **FIG. 1** is a block diagram of an embodiment of a transmitter system 110 and a receiver system 150, which are capable of implementing various signal processing techniques described herein.

[1028] At transmitter system 110, traffic data is provided from a data source 112 to a transmit (TX) data processor 114, which formats, codes, and interleaves the traffic data based on one or more coding schemes to provide coded data. The coded traffic data may then be multiplexed with pilot data using, for example, time division multiplex (TDM) or code division multiplex (CDM), in all or a subset of the data streams to be transmitted. The pilot data is typically a known data pattern processed in a known manner, if at all. The multiplexed pilot and coded traffic data is interleaved and then modulated (i.e., symbol mapped) based on one or more modulation schemes to provide modulation symbols. In an embodiment, TX data processor 114 provides one modulation symbol stream for each spatial subchannel used for data transmission. The data rate, coding, interleaving, and modulation for each modulation symbol stream may be determined by controls provided by a controller 130.

[1029] The modulation symbols are then provided to a TX MIMO processor 120 and further processed. In a specific embodiment, the processing by TX MIMO processor 120 includes (1) determining an estimated channel frequency response matrix for the MIMO channel, (2) decomposing this matrix to determine the eigenmodes of the MIMO channel and to derive a set of "steering" vectors for the transmitter, one vector for the modulation symbol stream to be transmitted on each spatial subchannel, (3) deriving a transmit spatio-temporal pulse-shaping matrix based on the steering vectors and a weighting matrix indicative of the transmit powers assigned to the frequency bins of the eigenmodes, and (4) preconditioning the modulation symbols with the pulse-shaping matrix to provide preconditioned modulation symbols. The processing by TX MIMO processor 120 is described in further detail below. Up to N_T streams of preconditioned symbols are then provided to transmitters (TMTR) 122a through 122t.

[1030] Each transmitter 122 converts a respective preconditioned symbol stream into one or more analog signals and further conditions (e.g., amplifies, filters, and frequency upconverts) the analog signals to generate a modulated signal suitable for

transmission over the MIMO channel. The modulated signal from each transmitter 122 is then transmitted via a respective antenna 124 to the receiver system.

[1031] At receiver system 150, the transmitted modulated signals are received by N_R antennas 152a through 152r, and the received signal from each antenna 152 is provided to a respective receiver (RCVR) 154. Each receiver 154 conditions (e.g., filters, amplifies, and frequency downconverts) the received signal, digitizes the conditioned signal to provide a stream of samples, and further processes the sample stream to provide a stream of received symbols. An RX MIMO processor 160 then receives and processes the N_R received symbol streams to provide N_T streams of recovered symbols, which are estimates of the modulation symbols transmitted from the transmitter system. In an embodiment, the processing by RX MIMO processor 160 may include (1) determining the estimated channel frequency response matrix for the MIMO channel, (2) decomposing this matrix to derive a set of steering vectors for the receiver, (3) deriving a receive spatio-temporal pulse-shaping matrix based on the steering vectors, and (4) conditioning the received symbols with the pulse-shaping matrix to provide the recovered symbols. The processing by RX MIMO processor 160 is described in further detail below.

[1032] A receive (RX) data processor 162 then demodulates, deinterleaves, and decodes the recovered symbols to provide decoded data, which is an estimate of the transmitted traffic data. The processing by RX MIMO processor 160 and RX data processor 162 is complementary to that performed by TX MIMO processor 120 and TX data processor 114, respectively, at transmitter system 110.

[1033] RX MIMO processor 160 may further derive channel impulse responses for the MIMO channel, received noise power and/or signal-to-noise-and-interference ratios (SNRs) for the spatial subchannels, and so on. RX MIMO processor 160 would then provide these quantities to a controller 170. RX data processor 162 may also provide the status of each received packet or frame, one or more other performance metrics indicative of the decoded results, and possibly other information. Controller 170 then derives channel state information (CSI), which may comprise all or some of the information received from RX MIMO processor 160 and RX data processor 162. The CSI is processed by a TX data processor 178, modulated by a modulator 180, conditioned by transmitters 154a through 154r, and sent back to transmitter system 110.

[1034] At transmitter system 110, the modulated signals from receiver system 150 are received by antennas 124, conditioned by receivers 122, and demodulated by a demodulator 140 to recover the CSI transmitted by the receiver system. The CSI is then provided to controller 130 and used to generate various controls for TX data processor 114 and TX MIMO processor 120.

[1035] Controllers 130 and 170 direct the operation at the transmitter and receiver systems, respectively. Memories 132 and 172 provide storage for program codes and data used by controllers 130 and 170, respectively.

[1036] Techniques are provided herein for achieving high performance (e.g., high overall system throughput) via a time-domain implementation that uses frequency-domain channel eigen-decomposition, channel inversion, and (optionally) water-pouring results to derive time-domain pulse-shaping and beam-steering solutions for the transmitter and receiver.

[1037] Channel eigen-decomposition is performed at the transmitter to determine the eigenmodes of the MIMO channel and to derive a first set of steering vectors, which are used to precondition the modulation symbols. Channel eigen-decomposition is also performed at the receiver to derive a second set of steering vectors, which are used to condition the received symbols such that orthogonal symbol streams are recovered at the receiver. The preconditioning at the transmitter and the conditioning at the receiver orthogonalize the symbol streams transmitted over the MIMO channel.

[1038] Channel inversion is performed at the transmitter to flatten the frequency response of each eigenmode (or spatial subchannel) used for data transmission. As noted above, frequency selective fading causes intersymbol interference (ISI), which can degrade performance by impacting the ability to correctly detect the received symbols at the receiver. Conventionally, the frequency selective fading may be compensated for at the receiver by performing equalization on the received symbol streams. However, equalization increases the complexity of the receiver processing. With the inventive techniques, the channel inversion is performed at the transmitter to account for the frequency selective fading and to mitigate the need for equalization at the receiver.

[1039] Water-pouring (or water-filling) analysis is used to more optimally allocate the total available transmit power in the MIMO system to the eigenmodes such that high performance is achieved. The transmit power allocated to each eigenmode may then

determine the data rate and the coding and modulation scheme to be used for the eigenmode.

[1040] These various processing techniques are described in further detail below.

[1041] The techniques described herein provide several potential advantages. First, with time-domain eigenmode decomposition, the maximum number of eigenmodes with different SNRs is given by $\min(N_T, N_R)$. If one independent data stream is transmitted on each eigenmode and each data stream is independently processed, then the maximum number of different coding/modulation schemes is also given by $\min(N_T, N_R)$. It is also possible to make the received SNRs for the data streams essentially the same, thereby further simplifying the coding/modulation. The techniques described herein can thus greatly simplify the coding/modulation for a data transmission by avoiding the per-bin bit allocation required to approach channel capacity in MIMO-OFDM systems that utilize frequency-domain water-pouring.

[1042] Second, the channel inversion at the transmitter results in recovered symbol streams at the receiver that do not require equalization. This then reduces the complexity of the receiver processing. In contrast, other wide-band time-domain techniques typically require complicated space-time equalization to recover the symbol streams.

[1043] Third, the time-domain signaling techniques described herein can more easily integrate the channel/pilot structures of various CDMA standards, which are also based on time-domain signaling. Implementation of the channel/pilot structures may be more complicated in OFDM-based systems that perform frequency-domain signaling.

[1044] **FIG. 2** is a block diagram of an embodiment of a transmitter unit 200, which is capable of implementing various processing techniques described herein. Transmitter unit 200 is an embodiment of the transmitter portion of transmitter system 110 in FIG. 1. Transmitter unit 200 includes (1) a TX data processor 114a that receives and processes traffic and pilot data to provide N_T modulation symbol streams and (2) a TX MIMO processor 120a that preconditions the modulation symbol streams to provide N_T preconditioned symbol streams. TX data processor 114a and TX MIMO processor 120a are one embodiment of TX data processor 114 and TX MIMO processor 120, respectively, in FIG. 1.

[1045] In the specific embodiment shown in FIG. 2, TX data processor 114a includes an encoder 212, a channel interleaver 214, and a symbol mapping element 216. Encoder 212 receives and codes the traffic data (i.e., the information bits, d_i) in accordance with one or more coding schemes to provide coded bits. The coding increases the reliability of the data transmission. In an embodiment, a separate coding scheme may be used for the information bits for each eigenmode (or spatial subchannel) selected for use for data transmission. In alternative embodiments, a separate coding scheme may be used for each subset of spatial subchannels, or a common coding scheme may be used for all spatial subchannels. The coding scheme(s) to be used are determined by controls from controller 130 and may be selected based on the CSI received from the receiver system. Each selected coding scheme may include any combination of cyclic redundancy check (CRC), convolutional coding, Turbo coding, block coding, and other coding, or no coding at all.

[1046] Channel interleaver 214 interleaves the coded bits based on one or more interleaving schemes. Typically, each selected coding scheme is associated with a corresponding interleaving scheme. The interleaving provides time diversity for the coded bits, permits the data to be transmitted based on an average SNR of each spatial subchannel used for the data transmission, combats fading, and further removes correlation between coded bits used to form each modulation symbol.

[1047] Symbol mapping element 216 then receives and multiplexes pilot data with the interleaved data and further maps the multiplexed data in accordance with one or more modulation schemes to provide modulation symbols. A separate modulation scheme may be used for each spatial subchannel selected for use, or for each subset of spatial subchannels. Alternatively, a common modulation scheme may be used for all selected spatial subchannels.

[1048] The symbol mapping for each spatial subchannel may be achieved by grouping sets of bits to form data symbols (each of which may be a non-binary value) and mapping each data symbol to a point in a signal constellation corresponding to the modulation scheme selected for use for that spatial subchannel. The selected modulation scheme may be QPSK, M-PSK, M-QAM, or some other scheme. Each mapped signal point is a complex value and corresponds to a modulation symbol. Symbol mapping element 216 provides a vector of modulation symbols for each symbol

period, with the number of modulation symbols in each vector corresponding to the number of spatial subchannels selected for use for that symbol period. Symbol mapping element 216 thus provides up to N_T modulation symbol streams. These streams collectively form a sequence of vectors, with are also referred to as the modulation symbol vectors, $\underline{s}(n)$, with each such vector including up to N_S modulation symbols to be transmitted on up to N_S spatial subchannels for the n-th symbol period.

[1049] Within TX MIMO processor 120a, the response of the MIMO channel is estimated and used to precondition the modulation symbols prior to transmission to the receiver system. In a frequency division duplexed (FDD) system, the downlink and uplink are allocated different frequency bands, and the channel responses for the downlink and uplink may not be correlated to a sufficient degree. For the FDD system, the channel response may be estimated at the receiver and sent back to the transmitter. In a time division duplexed (TDD) system, the downlink and uplink share the same frequency band in a time division multiplexed manner, and a high degree of correlation may exist between the downlink and uplink channel responses. For the TDD system, the transmitter system may estimate the uplink channel response (e.g., based on the pilot transmitted by the receiver system on the uplink) and may then derive the downlink channel response by accounting for any differences such as those between the transmit and receive antenna array manifolds.

[1050] In an embodiment, the channel response estimates are provided to TX MIMO processor 120a as a sequence of $N_R \times N_T$ matrices, $\underline{\hat{\mathcal{H}}}(n)$, of time-domain samples. This sequence of matrices is collectively referred to as a channel impulse response matrix, $\underline{\underline{\hat{\mathcal{H}}}}$. The (i, j)-th element, $\underline{\hat{h}}_{i,j}$, of the estimated channel impulse response matrix, $\underline{\underline{\hat{\mathcal{H}}}}$, for $i = (1, 2, \ldots, N_R)$ and $j = (1, 2, \ldots, N_T)$, is a sequence of samples that represents the sampled impulse response of the propagation path from the j-th transmit antenna to the i-th receive antenna.

[1051] Within TX MIMO processor 120a, a fast Fourier transformer 222 receives the estimated channel impulse response matrix, $\underline{\underline{\hat{\mathcal{H}}}}$ (e.g., from the receiver system) and derives the corresponding estimated channel frequency response matrix, $\underline{\underline{\hat{\mathbf{H}}}}$, by performing a fast Fourier transform (FFT) on the matrix $\underline{\underline{\hat{\mathcal{H}}}}$ (i.e., $\underline{\underline{\hat{\mathbf{H}}}} = \text{FFT}[\underline{\underline{\hat{\mathcal{H}}}}]$). This

may be achieved by performing an N_F-point FFT on a sequence of N_F samples for each element of $\underline{\underline{\hat{\mathcal{H}}}}$ to derive a set of N_F coefficients for the corresponding element of $\underline{\underline{\hat{\mathbf{H}}}}$, where N_F corresponds to the number of frequency bins for the FFT (i.e., the length of the FFT). The $N_R \cdot N_T$ elements of $\underline{\underline{\hat{\mathbf{H}}}}$ are thus $N_R \cdot N_T$ sets of coefficients representing the frequency responses of the propagation paths between the N_T transmit antennas and N_R receive antennas. Each element of $\underline{\underline{\hat{\mathbf{H}}}}$ is the FFT of the corresponding element of $\underline{\underline{\hat{\mathcal{H}}}}$. The estimated channel frequency response matrix, $\underline{\underline{\hat{\mathbf{H}}}}$, may also be viewed as comprising a set of N_F matrices, $\hat{\underline{\mathbf{H}}}(k)$ for $k = (0, 1, \ldots, N_F - 1)$.

Channel Eigen-Decomposition

[1052] A unit 224 then performs eigen-decomposition of the MIMO channel used for data transmission. In one embodiment for performing channel eigen-decomposition, unit 224 computes the singular value decomposition (SVD) of the estimated channel frequency response matrix, $\underline{\underline{\hat{\mathbf{H}}}}$. In an embodiment, the singular value decomposition is performed for each matrix $\hat{\underline{\mathbf{H}}}(k)$, for $k = (0, 1, \ldots, N_F - 1)$. The singular value decomposition of matrix $\hat{\underline{\mathbf{H}}}(k)$ for frequency bin k (or frequency f_k) may be expressed as:

$$\hat{\underline{\mathbf{H}}}(k) = \underline{\mathbf{U}}(k)\underline{\Lambda}(k)\underline{\mathbf{V}}^H(k) \quad , \qquad \text{Eq (1)}$$

where $\underline{\mathbf{U}}(k)$ is an $N_R \times N_R$ unitary matrix (i.e., $\underline{\mathbf{U}}^H \underline{\mathbf{U}} = \underline{\mathbf{I}}$, where $\underline{\mathbf{I}}$ is the identity matrix with ones along the diagonal and zeros everywhere else);

$\underline{\Lambda}(k)$ is an $N_R \times N_T$ diagonal matrix of singular values of $\hat{\underline{\mathbf{H}}}(k)$; and

$\underline{\mathbf{V}}(k)$ is an $N_T \times N_T$ unitary matrix.

The diagonal matrix $\underline{\Lambda}(k)$ contains non-negative real values along the diagonal (i.e., $\underline{\Lambda}(k) = \text{diag}\,(\lambda_1(k), \lambda_2(k), \ldots, \lambda_{N_T}(k))$) and zeros elsewhere. The $\lambda_i(k)$, for $i = (1, 2, \ldots, N_T)$, are referred to as the singular values of the matrix $\hat{\underline{\mathbf{H}}}(k)$. The singular value decomposition is a matrix operation known in the art and described in various references. One such reference is a book by Gilbert Strang entitled "Linear

Algebra and Its Applications," Second Edition, Academic Press, 1980, which is incorporated herein by reference.

[1053] The result of the singular value decomposition is three sets of N_F matrices, $\underline{\underline{\mathbf{U}}}$, $\underline{\underline{\Lambda}}$, and $\underline{\underline{\mathbf{V}}}^H$, where $\underline{\underline{\mathbf{U}}} = [\underline{\mathbf{U}}(0) \ \ldots \ \underline{\mathbf{U}}(k) \ \ldots \ \underline{\mathbf{U}}(N_F - 1)]$, and so on. For each value of k, $\underline{\mathbf{U}}(k)$ is the $N_R \times N_R$ unitary matrix of left eigen-vectors of $\hat{\underline{\mathbf{H}}}(k)$, $\underline{\mathbf{V}}(k)$ is the $N_T \times N_T$ unitary matrix of right eigen-vectors of $\hat{\underline{\mathbf{H}}}(k)$, and $\underline{\Lambda}(k)$ is the $N_R \times N_T$ diagonal matrix of singular values of $\hat{\underline{\mathbf{H}}}(k)$.

[1054] In another embodiment for performing channel eigen-decomposition, unit 224 first obtains a square matrix $\underline{\mathbf{R}}(k)$ as $\underline{\mathbf{R}}(k) = \hat{\underline{\mathbf{H}}}^H(k)\hat{\underline{\mathbf{H}}}(k)$. The eigenvalues of the square matrix $\underline{\mathbf{R}}(k)$ would then be the squares of the singular values of the matrix $\hat{\underline{\mathbf{H}}}(k)$, and the eigen-vectors of $\underline{\mathbf{R}}(k)$ would be the right eigen-vectors of $\hat{\underline{\mathbf{H}}}(k)$, or $\underline{\mathbf{V}}(k)$. The decomposition of $\underline{\mathbf{R}}(k)$ to obtain the eigenvalues and eigen-vectors is known in the art and not described herein. Similarly, another square matrix $\underline{\mathbf{R}}'(k)$ may be obtained as $\underline{\mathbf{R}}'(k) = \hat{\underline{\mathbf{H}}}(k)\hat{\underline{\mathbf{H}}}^H(k)$. The eigenvalues of this square matrix $\underline{\mathbf{R}}'(k)$ would also be the squares of the singular values of $\hat{\underline{\mathbf{H}}}(k)$, and the eigen-vectors of $\underline{\mathbf{R}}'(k)$ would be the left eigen-vectors of $\hat{\underline{\mathbf{H}}}(k)$, or $\underline{\mathbf{U}}(k)$.

[1055] The channel eigen-decomposition is used to decompose the MIMO channel into its eigenmodes, at frequency f_k, for each value of k where $k = (0, 1, \ldots, N_F - 1)$. The rank $r(k)$ of $\hat{\underline{\mathbf{H}}}(k)$ corresponds to the number of eigenmodes for the MIMO channel at frequency f_k, which corresponds to the number of independent channels (i.e., the number of spatial subchannels) available in frequency bin k.

[1056] As described in further detail below, the columns of $\underline{\mathbf{V}}(k)$ are the steering vectors associated with frequency f_k to be used at the transmitter for the elements of the modulation symbol vectors, $\underline{\mathbf{s}}(n)$. Correspondingly, the columns of $\underline{\mathbf{U}}(k)$ are the steering vectors associated with frequency f_k to be used at the receiver for the elements of the received symbol vectors, $\underline{\mathbf{r}}(n)$. The matrices $\underline{\mathbf{U}}(k)$ and $\underline{\mathbf{V}}(k)$, for $k = (0, 1, \ldots, N_F - 1)$, are used to orthogonalize the symbol streams transmitted on the

eigenmodes at each frequency f_k. When these matrices are used to precondition the modulation symbol streams at the transmitter and to condition the received symbol streams at the receiver, either in the frequency domain or the time domain, the result is the overall orthogonalization of the symbol streams. This then allows for separate coding/modulation per eigenmode (as opposed to per bin), which can greatly simplify the processing at the transmitter and receiver.

[1057] The elements along the diagonal of $\underline{\Lambda}(k)$ are $\lambda_{ii}(k)$, for $i = \{1, 2, \ldots, r(k)\}$, where $r(k)$ is the rank of $\hat{\underline{\mathbf{H}}}(k)$. The columns of $\underline{\mathbf{U}}(k)$ and $\underline{\mathbf{V}}(k)$, $\underline{\mathbf{u}}_i(k)$ and $\underline{\mathbf{v}}_i(k)$, respectively, are solutions to the eigen equation, which may be expressed as:

$$\hat{\underline{\mathbf{H}}}(k)\underline{\mathbf{v}}_i(k) = \lambda_{ii}\underline{\mathbf{u}}_i(k) \quad . \qquad \text{Eq (2)}$$

[1058] The three sets of matrices, $\underline{\mathbf{U}}(k)$, $\underline{\Lambda}(k)$, and $\underline{\mathbf{V}}(k)$, for $k = (0, 1, \ldots, N_F - 1)$, may be provided in two forms - a "sorted" form and a "random-ordered" form. In the sorted form, the diagonal elements of each matrix $\underline{\Lambda}(k)$ are sorted in decreasing order so that $\lambda_{11}(k) \geq \lambda_{22}(k) \geq \ldots \geq \lambda_{rr}(k)$, and their eigen-vectors are arranged in corresponding order in $\underline{\mathbf{U}}(k)$ and $\underline{\mathbf{V}}(k)$. The sorted form is indicated by the subscript *s*, i.e., $\underline{\mathbf{U}}_s(k)$, $\underline{\Lambda}_s(k)$, and $\underline{\mathbf{V}}_s(k)$, for $k = (0, 1, \ldots, N_F - 1)$.

[1059] In the random-ordered form, the ordering of the singular values and eigen-vectors may be random and further independent of frequency. The random form is indicated by the subscript *r*. The particular form selected for use, sorted or random-ordered, influences the selection of the eigenmodes for use for data transmission and the coding and modulation scheme to be used for each selected eigenmode.

[1060] A weight computation unit 230 receives the set of diagonal matrices, $\underline{\underline{\Lambda}}$, which contains a set of singular values (i.e., $\lambda_{11}(k), \lambda_{22}(k), \ldots, \lambda_{rr}(k)$) for each frequency bin. Weight computation unit 230 then derives a set of weighting matrices, $\underline{\underline{\mathbf{W}}}$, where $\underline{\underline{\mathbf{W}}} = [\underline{\mathbf{W}}(0) \ \ldots \ \underline{\mathbf{W}}(k) \ \ldots \ \underline{\mathbf{W}}(N_F - 1)]$. The weighting matrices are used to scale the modulation symbol vectors, $\underline{\mathbf{s}}(n)$, in either the time or frequency domain, as described below.

[1061] Weight computation unit 230 includes a channel inversion unit 232 and (optionally) a water-pouring analysis unit 234. Channel inversion unit 232 derives a set of weights, $\underline{\mathbf{w}}_{ii}$, for each eigenmode, which is used to combat the frequency selective fading on the eigenmode. Water-pouring analysis unit 234 derives a set of scaling values, $\underline{\mathbf{b}}$, for the eigenmodes of the MIMO channel. These scaling values are indicative of the transmit powers allocated to the eigenmodes. Channel inversion and water-pouring are described in further detail below.

Channel Inversion

[1062] **FIG. 3A** is a diagram that graphically illustrates the derivation of the set of weights, $\underline{\mathbf{w}}_{ii}$, used to invert the frequency response of each eigenmode. The set of diagonal matrices, $\underline{\Lambda}(k)$ for $k = (0, 1, \ldots, N_F - 1)$, is shown arranged in order along an axis 310 that represents the frequency dimension. The singular values, $\lambda_{ii}(k)$ for $i = (1, 2, \ldots, N_S)$, of each matrix $\underline{\Lambda}(k)$ are located along the diagonal of the matrix. Axis 312 may thus be viewed as representing the spatial dimension. Each eigenmode of the MIMO channel is associated with a set of elements, $\{\lambda_{ii}(k)\}$ for $k = (0, 1, \ldots, N_F - 1)$, that is indicative of the frequency response of that eigenmode. The set of elements $\{\lambda_{ii}(k)\}$ for each eigenmode is shown by the shaded boxes along a dashed line 314. For each eigenmode that experiences frequency selective fading, the elements $\{\lambda_{ii}(k)\}$ for the eigenmode may be different for different values of k.

[1063] Since frequency selective fading causes ISI, the deleterious effects of ISI may be mitigated by "inverting" each eigenmode such that it appears flat in frequency at the receiver. The channel inversion may be achieved by deriving a set of weights, $\{w_{ii}(k)\}$ for $k = (0, 1, \ldots, N_F - 1)$, for each eigenmode such that the product of the weights and the corresponding eigenvalues (i.e., the squares of the diagonal elements) are approximately constant for all values of k, which may be expressed as $w_{ii}(k) \cdot \lambda_{ii}^2(k) = a_i$, for $k = (0, 1, \ldots, N_F - 1)$.

[1064] For eigenmode i, the set of weights for the N_F frequency bins, $\underline{\mathbf{w}}_{ii} = [w_{ii}(0) \ldots w_{ii}(k) \ldots w_{ii}(N_F - 1)]^T$, used to invert the channel may be derived as:

$$w_{ii}(k)=\frac{a_i}{\lambda_{ii}^2(k)}\ ,\qquad \text{for } k=(0,\ 1,\ \ldots,N_F-1)\,, \qquad \text{Eq (3)}$$

where a_i is a normalization factor that may be expressed as:

$$a_i=\frac{\sum_{k=0}^{N_F-1}\lambda_{ii}^2(k)}{\sum_{k=0}^{N_F-1}\frac{1}{\lambda_{ii}^2(k)}}\ . \qquad \text{Eq (4)}$$

As shown in equation (4), a normalization factor a_i is determined for each eigenmode based on the set of eigenvalues (i.e., the squared singular values), $\{\lambda_{ii}^2(k)\}$ for $k=(0,\ 1,\ \ldots,N_F-1)$, associated with that eigenmode. The normalization factor a_i is defined such that $\sum_{k=0}^{N_F-1}w_{ii}(k)=\sum_{k=0}^{N_F-1}\lambda_{ii}^2(k)$.

[1065] **FIG. 3B** is a diagram that graphically illustrates the relationship between the set of weights for a given eigenmode and the set of eigenvalues for that eigenmode. For eigenmode *i*, the weight $w_{ii}(k)$ for each frequency bin is inversely related to the eigenvalue $\lambda_{ii}^2(k)$ for that bin, as shown in equation (3). To flatten the spatial subchannel and minimize or reduce ISI, it is undesirable to selectively eliminate transmit power on any frequency bin. The set of N_F weights for each eigenmode is used to scale the modulation symbols, $\underline{s}(n)$, in the frequency or time domain, prior to transmission on the eigenmode.

[1066] For the sorted order form, the singular values $\lambda_{ii}(k)$, for $i=(1,\ 2,\ \ldots,N_S)$, for each matrix $\underline{\Lambda}(k)$ are sorted such that the diagonal elements of $\underline{\Lambda}(k)$ with smaller indices are generally larger. Eigenmode 0 (which is often referred to as the principle eigenmode) would then be associated with the largest singular value in each of the N_F diagonal matrices, $\underline{\Lambda}(k)$, eigenmode 1 would then be associated with the second largest singular value in each of the N_F diagonal matrices, and so on. Thus, even though the channel inversion is performed over all N_F frequency bins for each eigenmode, the eigenmodes with lower indices are not likely to have too many bad bins (if any). Thus,

at least for eigenmodes with lower indices, excessive transmit power is not used for bad bins.

[1067] The channel inversion may be performed in various manners to invert the MIMO channel, and this is within the scope of the invention. In one embodiment, the channel inversion is performed for each eigenmode selected for use. In another embodiment, the channel inversion may be performed for some eigenmodes but not others. For example, the channel inversion may be performed for each eigenmode determined to induce excessive ISI. The channel inversion may also be dynamically performed for some or all eigenmodes selected for use, for example, when the MIMO channel is determined to be frequency selective (e.g., based on some defined criteria).

[1068] Channel inversion is described in further detail in U.S. Patent Application Serial No. 09/860,274, filed May 17, 2001, U.S. Patent Application Serial No. 09/881,610, filed June 14, 2001, and U.S. Patent Application Serial No. 09/892,379, filed June 26, 2001, all three entitled "Method and Apparatus for Processing Data for Transmission in a Multi-Channel Communication System Using Selective Channel Inversion," assigned to the assignee of the present application and incorporated herein by reference.

Water-Pouring

[1069] In an embodiment, water-pouring analysis is performed (if at all) across the spatial dimension such that more transmit power is allocated to eigenmodes with better transmission capabilities. The water-pouring power allocation is analogous to pouring a fixed amount of water into a vessel with an irregular bottom, where each eigenmode corresponds to a point on the bottom of the vessel, and the elevation of the bottom at any given point corresponds to the inverse of the SNR associated with that eigenmode. A low elevation thus corresponds to a high SNR and, conversely, a high elevation corresponds to a low SNR. The total available transmit power, P_{total}, is then "poured" into the vessel such that the lower points in the vessel (i.e., those with higher SNRs) are filled first, and the higher points (i.e., those with lower SNRs) are filled later. A constant P_{set} is indicative of the water surface level for the vessel after all of the total available transmit power has been poured. This constant may be estimated initially based on various system parameters. The power allocation is dependent on the total available transmit power and the depth of the vessel over the bottom surface. The

points with elevations above the water surface level are not filled (i.e., eigenmodes with SNRs below a particular value are not used for data transmission).

[1070] In an embodiment, the water-pouring is not performed across the frequency dimension because this tends to exaggerate the frequency selectivity of the eigenmodes created by the channel eigenmode decomposition described above. The water-pouring may be performed such that all eigenmodes are used for data transmission, or only a subset of the eigenmodes is used (with bad eigenmodes being discarded). It can be shown that water-pouring across the eigenmodes, when used in conjunction with the channel inversion with the singular values sorted in descending order, can provide near-optimum performance while mitigating the need for equalization at the receiver.

[1071] The water-pouring may be performed by water-pouring analysis unit 234 as follows. Initially, the total power in each eigenmode is determined as:

$$P_{i,\lambda} = \sum_{k=0}^{N_F-1} \lambda_{ii}^2(k) \ . \qquad \text{Eq (5)}$$

[1072] The SNR for each eigenmode may then be determined as:

$$\text{SNR}_i = \frac{P_{i,\lambda}}{\sigma^2} \ , \qquad \text{Eq (6)}$$

where σ^2 is the received noise variance, which may also be denoted as the received noise power N_0. The received noise power corresponds to the noise power on the recovered symbols at the receiver, and is a parameter that may be provided by the receiver to the transmitter as part of the reported CSI.

[1073] The transmit power, P_i, to be allocated to each eigenmode may then be determined as:

$$P_i = \max\left[\left(P_{set} - \frac{1}{\text{SNR}_i}\right),\ 0\right] \ , \text{ and} \qquad \text{Eq (7a)}$$

$$P_{total} \geq \sum_{i=1}^{N_S} P_i \ , \qquad \text{Eq (7b)}$$

where P_{set} is a constant that may be derived from various system parameters, and P_{total} is the total transmit power available for allocation to the eigenmodes.

[1074] As shown in equation (7a), each eigenmode of sufficient quality is allocated transmit power of $\left(P_{set} - \frac{1}{\mathrm{SNR}_i}\right)$. Thus, eigenmodes that achieve better SNRs are allocated more transmit powers. The constant P_{set} determines the amounts of transmit power to allocate to the better eigenmodes. This then indirectly determines which eigenmodes get selected for use since the total available transmit power is limited and the power allocation is constrained by equation (7b).

[1075] Water-pouring analysis unit 234 thus receives the set of diagonal matrices, $\underline{\underline{\Lambda}}$, and the received noise power, σ^2. The matrices $\underline{\underline{\Lambda}}$ are then used in conjunction with the received noise power to derive a vector of scaling values, $\underline{\mathbf{b}} = [b_0 \ \dots \ b_i \ \dots \ b_{N_S}]^T$, where $b_i = P_i$ for $i = (1, 2, \dots, N_S)$. The P_i are the solutions to the water-pouring equations (7a) and (7b). The scaling values in $\underline{\mathbf{b}}$ are indicative of the transmit powers allocated to the N_S eigenmodes, where zero or more eigenmodes may be allocated no transmit power.

[1076] **FIG. 4** is a flow diagram of an embodiment of a process 400 for allocating the total available transmit power to a set of eigenmodes. Process 400, which is one specific water-pouring implementation, determines the transmit powers, P_i, for $i \in I$, to be allocated to the eigenmodes in set I, given the total transmit power, P_{total}, available at the transmitter, the set of eigenmode total powers, $P_{i,\lambda}$, and the received noise power, σ^2.

[1077] Initially, the variable n used to denote the iteration number is set to one (i.e., $n = 1$) (step 412). For the first iteration, set $I(n)$ is defined to include all of the eigenmodes for the MIMO channel, or $I(n) = \{1, 2, \dots, N_S\}$ (step 414). The cardinality (or length) of set $I(n)$ for the current iteration n is then determined as $L_I(n) = |I(n)|$, which is $L_I(n) = N_S$ for the first iteration (step 416).

[1078] The total effective power, $P_{eff}(n)$, to be distributed across the eigenmodes in set $I(n)$ is next determined (step 418). The total effective power is defined to be equal

to the total available transmit power, P_{total}, plus the sum of the inverse SNRs for the eigenmodes in set $I(n)$. This may be expressed as:

$$P_{eff}(n) = P_{total} + \sum_{i \in I(n)} \frac{\sigma^2}{P_{i,\lambda}} \quad . \qquad \text{Eq (8)}$$

[1079] The total available transmit power is then allocated to the eigenmodes in set $I(n)$. The index i used to iterate through the eigenmodes in set $I(n)$ is initialized to one (i.e., $i = 1$) (step 420). The amount of transmit power to allocate to eigenmode i is then determined (step 422) based on the following:

$$P_i(n) = \frac{P_{eff}(n)}{L_I(n)} - \frac{\sigma^2}{P_{i,\lambda}} \quad . \qquad \text{Eq (9)}$$

Each eigenmode in set $I(n)$ is allocated transmit power, P_i, in step 422. Steps 424 and 426 are part of a loop to allocate transmit power to each of the eigenmodes in set $I(n)$.

[1080] **FIG. 5A** graphically illustrates the total effective power, P_{eff}, for an example MIMO system with three eigenmodes. Each eigenmode has an inverse SNR equal to $\sigma^2 / \lambda_{ii}^2$, for $i = \{1, 2, 3\}$, which assumes a normalized transmit power of 1.0. The total transmit power available at the transmitter is $P_{total} = P_1 + P_2 + P_3$, and is represented by the shaded area in FIG. 5A. The total effective power is represented by the area in the shaded and unshaded regions in FIG. 5A.

[1081] For water-pouring, although the bottom of the vessel has an irregular surface, the water level at the top remains constant across the vessel. Likewise and as shown in FIG. 5A, after the total available transmit power, P_{total}, has been distributed to the eigenmodes, the final power level is constant across all eigenmodes. This final power level is determined by dividing $P_{eff}(n)$ by the number of eigenmodes in set $I(n)$, $L_I(n)$. The amount of power allocated to eigenmode i is then determined by subtracting the inverse SNR of that eigenmode, $\sigma^2 / \lambda_{ii}^2$, from the final power level, $P_{eff}(n) / L_I(n)$, as given by equation (9) and shown in FIG. 5A.

[1082] **FIG. 5B** shows a situation whereby the water-pouring power allocation results in an eigenmode receiving negative power. This occurs when the inverse SNR

of the eigenmode is above the final power level, which is expressed by the condition $(P_{eff}(n)/L_I(n)) < (\sigma^2/\lambda_{ii}^2)$.

[1083] Referring back to FIG. 4, at the end of the power allocation, a determination is made whether or not any eigenmode has been allocated negative power (i.e., $P_i < 0$) (step 428). If the answer is yes, then the process continues by removing from set $I(n)$ all eigenmodes that have been allocated negative powers (step 430). The index n is incremented by one (i.e., $n = n+1$) (step 432). The process then returns to step 416 to allocate the total available transmit power among the remaining eigenmodes in set $I(n)$. The process continues until all eigenmodes in set $I(n)$ have been allocated positive transmit powers, as determined in step 428. The eigenmodes not in set $I(n)$ are allocated zero power.

[1084] Water-pouring is also described by Robert G. Gallager, in "Information Theory and Reliable Communication," John Wiley and Sons, 1968, which is incorporated herein by reference. A specific algorithm for performing the basic water-pouring process for a MIMO-OFDM system is described in U.S. Patent Application Serial No. 09/978,337, entitled "Method and Apparatus for Determining Power Allocation in a MIMO Communication System," filed October 15, 2001. Water-pouring is also described in further detail in U.S. Patent Application Serial No. 10/056,275, entitled "Reallocation of Excess Power for Full Channel-State Information (CSI) Multiple-Input, Multiple-Output (MIMO) Systems," filed January 23, 2002. These applications are assigned to the assignee of the present application and incorporated herein by reference.

[1085] If water-pouring is performed to allocate the total available transmit power to the eigenmodes, then water-pouring analysis unit 234 provides a set of N_S scaling values, $\underline{\mathbf{b}} = \{b_0 \ \ldots \ b_i \ \ldots \ b_{N_S}\}$, for the N_S eigenmodes. Each scaling value is for a respective eigenmode and is used to scale the set of weights determined for that eigenmode.

[1086] For eigenmode i, a set of weights, $\underline{\hat{\mathbf{w}}}_{ii} = [\hat{w}_{ii}(0) \ \ldots \ \hat{w}_{ii}(k) \ \ldots \ \hat{w}_{ii}(N_F - 1)]^T$, used to invert the channel and scale the transmit power of the eigenmode may be derived as:

$$\hat{w}_{ii}(k) = \frac{a_i b_i}{\lambda_{ii}^2(k)} \quad , \qquad \text{for } k = (0,\ 1,\ \dots\ ,N_F - 1)\,, \qquad \text{Eq (10)}$$

where the normalization factor, a_i, and the scaling value, b_i, are derived as described above.

[1087] Weight computation unit 230 provides the set of weighting matrices, $\underline{\underline{\mathbf{W}}}$, which may be obtained using the weights $w_{ii}(k)$ or $\hat{w}_{ii}(k)$. Each weighting matrix, $\underline{\mathbf{W}}(k)$, is a diagonal matrix whose diagonal elements are composed of the weights derived above. In particular, if only channel inversion is performed, then each weighting matrix, $\underline{\mathbf{W}}(k)$, for $k = (0,\ 1,\ \dots\ ,N_F - 1)$, is defined as:

$$\underline{\mathbf{W}}(k) = \mathrm{diag}\,(w_{11}(k),\ w_{22}(k),\ \dots\ ,w_{N_S N_S}(k)) \quad , \qquad \text{Eq (11a)}$$

where $w_{ii}(k)$ is derived as shown in equation (3). And if both channel inversion and water-pouring are performed, then each weighting matrix, $\underline{\mathbf{W}}(k)$, for $k = (0,\ 1,\ \dots\ ,N_F - 1)$, is defined as:

$$\underline{\mathbf{W}}(k) = \mathrm{diag}\,(\hat{w}_{11}(k),\ \hat{w}_{22}(k),\ \dots\ ,\hat{w}_{N_S N_S}(k)) \quad , \qquad \text{Eq (11b)}$$

where $\hat{w}_{ii}(k)$ is derived as shown in equation (10).

[1088] Referring back to FIG. 2, a scaler/IFFT 236 receives (1) the set of unitary matrices, $\underline{\underline{\mathbf{V}}}$, which are the matrices of right eigen-vectors of $\underline{\underline{\hat{\mathbf{H}}}}$, and (2) the set of weighting matrices, $\underline{\underline{\mathbf{W}}}$, for all N_F frequency bins. Scaler/IFFT 236 then derives a spatio-temporal pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, for the transmitter based on the received matrices. Initially, the square root of each weighting matrix, $\underline{\mathbf{W}}(k)$, is computed to obtain a corresponding matrix, $\sqrt{\underline{\mathbf{W}}(k)}$, whose elements are the square roots of the elements of $\underline{\mathbf{W}}(k)$. The elements of the weighting matrices, $\underline{\mathbf{W}}(k)$ for $k = (0,\ 1,\ \dots\ ,N_F - 1)$, are related to the power of the eigenmodes. The square root then transforms the power to the equivalent signal scaling. For each frequency bin k, the product of the square-root weighting matrix, $\sqrt{\underline{\mathbf{W}}(k)}$, and the corresponding unitary

matrix, $\underline{\mathbf{V}}(k)$, is then computed to provide a product matrix, $\underline{\mathbf{V}}(k)\sqrt{\underline{\mathbf{W}}(k)}$. The set of product matrices, $\underline{\mathbf{V}}(k)\sqrt{\underline{\mathbf{W}}(k)}$ for $k = (0, 1, \ldots, N_F - 1)$, which is also denoted as $\underline{\underline{\mathbf{V}}}\sqrt{\underline{\underline{\mathbf{W}}}}$, defines the optimal or near-optimal spatio-spectral shaping to be applied to the modulation symbol vectors, $\underline{\mathbf{s}}(n)$.

[1089] An inverse FFT of $\underline{\underline{\mathbf{V}}}\sqrt{\underline{\underline{\mathbf{W}}}}$ is then computed to derive the spatio-temporal pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, for the transmitter, which may be expressed as:

$$\underline{\mathbf{P}}_{tx}(n) = \mathrm{IFFT}\,[\underline{\underline{\mathbf{V}}}\sqrt{\underline{\underline{\mathbf{W}}}}] \quad . \qquad \text{Eq (12)}$$

The pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, is an $N_T \times N_T$ matrix. Each element of $\underline{\mathbf{P}}_{tx}(n)$ is a set of N_F time-domain values, which is obtained by the inverse FFT of a set of values for the corresponding element of the product matrices, $\underline{\underline{\mathbf{V}}}\sqrt{\underline{\underline{\mathbf{W}}}}$. Each column of $\underline{\mathbf{P}}_{tx}(n)$ is a steering vector for a corresponding element of $\underline{\mathbf{s}}(n)$.

[1090] A convolver 240 receives and preconditions the modulation symbol vectors, $\underline{\mathbf{s}}(n)$, with the pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, to provide the transmitted symbol vectors, $\underline{\mathbf{x}}(n)$. In the time domain, the preconditioning is a convolution operation, and the convolution of $\underline{\mathbf{s}}(n)$ with $\underline{\mathbf{P}}_{tx}(n)$ may be expressed as:

$$\underline{\mathbf{x}}(n) = \sum_{\ell} \underline{\mathbf{P}}_{tx}(\ell)\underline{\mathbf{s}}(n-\ell) \quad . \qquad \text{Eq (13)}$$

The matrix convolution shown in equation (13) may be performed as follows. To derive the i-th element of the vector $\underline{\mathbf{x}}(n)$ for time n, $x_i(n)$, the inner product of the i-th row of the matrix $\underline{\mathbf{P}}_{tx}(\ell)$ with the vector $\underline{\mathbf{s}}(n-\ell)$ is formed for a number of delay indices (e.g., $0 \le \ell \le (N_F - 1)$), and the results are accumulated to derive the element $x_i(n)$. The preconditioned symbol streams transmitted on each transmit antenna (i.e., each element of $\underline{\mathbf{x}}(n)$ or $x_i(n)$) is thus formed as a weighted combination of the N_R modulation symbol streams, with the weighting determined by the appropriate column of the matrix $\underline{\mathbf{P}}_{tx}(n)$. The process is repeated such that each element of $\underline{\mathbf{x}}(n)$ is obtained from a respective column of the matrix $\underline{\mathbf{P}}_{tx}(n)$ and the vector $\underline{\mathbf{s}}(n)$.

[1091] Each element of $\underline{\mathbf{x}}(n)$ corresponds to a sequence of preconditioned symbols to be transmitted over a respective transmit antenna. The N_T preconditioned symbol sequences collectively form a sequence of vectors, which are also referred to as the transmitted symbol vectors, $\underline{\mathbf{x}}(n)$, with each such vector including up to N_T preconditioned symbols to be transmitted from up to N_T transmit antennas for the n-th symbol period. The N_T preconditioned symbol sequences are provided to transmitters 122a through 122t and processed to derive N_T modulated signals, which are then transmitted from antennas 124a through 124t, respectively.

[1092] The embodiment shown in FIG. 2 performs time-domain beam-steering of the modulation symbol vectors, $\underline{\mathbf{s}}(n)$. The beam-steering may also be performed in the frequency domain. This can be done using techniques, such as the overlap-add method, which are well-known in the digital signal processing field, for implementing finite-duration impulse response (FIR) filters in the frequency domain. In this case, the sequences that make up the elements of the matrix $\underline{\mathbf{P}}_{tx}(n)$ for $n = (0,\ 1,\ \ldots\ , N_F - 1)$ are each padded with $N_O - N_F$ zeros, resulting in a matrix of zero-padded sequences, $\underline{\mathbf{q}}_{tx}(n)$, for $n = (0,\ 1,\ \ldots\ , N_O - 1)$. An N_O-point fast Fourier transform (FFT) is then computed for each zero-padded sequence in the matrix $\underline{\mathbf{q}}_{tx}(n)$, resulting in a matrix $\underline{\mathbf{Q}}_{tx}(k)$ for $k = (0,\ 1,\ \ldots\ , N_O - 1)$.

[1093] Also, the sequences of modulation symbols that make up the elements of $\underline{\mathbf{s}}(n)$ are each broken up into subsequences of length $N_{SS} = N_O - N_F + 1$. Each subsequence is then padded with $N_F - 1$ zeros to provide a corresponding vector of length N_O. The sequences of $\underline{\mathbf{s}}(n)$ are thus processed to provide sequences of vectors of length N_O, $\tilde{\underline{\mathbf{s}}}_\ell(n)$, where the subscript ℓ is the index for the vectors that correspond to the zero-padded subsequences. An N_O-point fast Fourier transform is then computed for each of the zero-padded subsequences, resulting in a sequence of frequency-domain vectors, $\tilde{\underline{\mathbf{S}}}_\ell(k)$, for different values of ℓ. Each vector $\tilde{\underline{\mathbf{S}}}_\ell(k)$, for a given ℓ, includes a set of frequency-domain vectors of length N_O (i.e., for $k = (0,\ 1,\ \ldots\ , N_O - 1)$). The matrix $\underline{\mathbf{Q}}_{tx}(k)$ is then multiplied with the vector $\tilde{\underline{\mathbf{S}}}_\ell(k)$, for each value of ℓ, where the pre-multiplication is performed for each value of k, i.e., for $k = (0,\ 1,\ \ldots\ , N_O - 1)$.

The inverse FFTs are then computed for the matrix-vector product $\underline{\underline{\mathbf{Q}}}_{tx}(k)\underline{\tilde{\mathbf{S}}}_{\ell}(k)$ to provide a set of time-domain subsequences of length N_O. The resulting subsequences are then reassembled, according to the overlap-add method, or similar means, as is well-known in the art, to form the desired output sequences.

[1094] **FIG. 6** is a flow diagram of an embodiment of a process 600 that may be performed at the transmitter unit to implement the various transmit processing techniques described herein. Initially, data to be transmitted (i.e., the information bits) is processed in accordance with a particular processing scheme to provide a number of streams of modulation symbols (step 612). As noted above, the processing scheme may include one or more coding schemes and one or more modulation schemes (e.g., a separate coding and modulation scheme for each modulation symbol stream).

[1095] An estimated channel response matrix for the MIMO channel is then obtained (step 614). This matrix may be the estimated channel impulse response matrix, $\underline{\underline{\hat{\mathcal{H}}}}$, or the estimated channel frequency response matrix, $\underline{\underline{\hat{\mathbf{H}}}}$, which may be provided to the transmitter from the receiver. The estimated channel response matrix is then decomposed (e.g., using channel eigen-decomposition) to obtain a set of matrices of right eigen-vectors, $\underline{\underline{\mathbf{V}}}$, and a set of matrices of singular values, $\underline{\underline{\Lambda}}$ (step 616).

[1096] A number of sets of weights, $\underline{\mathbf{w}}_{ii}$, are then derived based on the matrices of singular values (step 618). One set of weight may be derived for each eigenmode used for data transmission. These weights are used to reduce or minimize intersymbol interference at the receiver by inverting the frequency response of each eigenmode selected for use.

[1097] A set of scaling values, $\underline{\mathbf{b}}$, may also be derived based on the matrices of singular values (step 620). Step 620 is optional, as indicated by the dashed box for step 620 in FIG. 6. The scaling values may be derived using water-pouring analysis and are used to adjust the transmit powers of the selected eigenmodes.

[1098] A pulse-shaping matrix, $\underline{\mathbf{P}}_{tx}(n)$, is then derived based on the matrices of right eigen-vectors, $\underline{\underline{\mathbf{V}}}$, the sets of weights, $\underline{\mathbf{w}}_{ii}$, and (if available) the set of scaling values, $\underline{\mathbf{b}}$ (step 622). The streams of modulation symbols are then preconditioned (in either the time domain or frequency domain) based on the pulse-shaping matrix to

provide a number of streams of preconditioned symbols, $\underline{\mathbf{x}}(n)$, to be transmitted over the MIMO channel (step 624).

[1099] Time-domain transmit processing with channel eigenmode decomposition and water-pouring is described in further detail in U.S. Patent Application Serial No. 10/017,038, entitled "Time-Domain Transmit and Receive Processing with Channel Eigen-mode Decomposition for MIMO Systems," filed December 7, 2001, which is assigned to the assignee of the present application and incorporated herein by reference.

[1100] **FIG. 7** is a block diagram of an embodiment of a receiver unit 700 capable of implementing various processing techniques described herein. Receiver unit 700 is an embodiment of the receiver portion of receiver system 150 in FIG. 1. Receiver unit 700 includes (1) a RX MIMO processor 160a that processes N_R received symbol streams to provide N_T recovered symbol streams and (2) a RX data processor 162a that demodulates, deinterleaves, and decodes the recovered symbols to provide decoded bits. RX MIMO processor 160a and RX data processor 162a are one embodiment of RX MIMO processor 160 and RX data processor 162, respectively, in FIG. 1.

[1101] Referring back to FIG. 1, the transmitted signals from N_T transmit antennas are received by each of N_R antennas 152a through 152r. The received signal from each antenna is routed to a respective receiver 154, which is also referred to as a front-end processor. Each receiver 154 conditions (e.g., filters, amplifies, and frequency downconverts) a respective received signal, and further digitizes the conditioned signal to provide ADC samples. Each receiver 154 may further data demodulate the ADC samples with a recovered pilot to provide a respective stream of received symbols. Receivers 154a through 154r thus provide N_R received symbol streams. These streams collectively form a sequence of vectors, which are also referred to as the received symbol vectors, $\underline{\mathbf{r}}(n)$, with each such vector including N_R received symbols from the N_R receivers 154 for the n-th symbol period. The received symbol vectors, $\underline{\mathbf{r}}(n)$, are then provided to RX MIMO processor 160a.

[1102] Within RX MIMO processor 160a, a channel estimator 712 receives the vectors $\underline{\mathbf{r}}(n)$ and derives an estimated channel impulse response matrix, $\hat{\underline{\underline{\mathcal{H}}}}$, which may be sent back to the transmitter system and used in the transmit processing. An FFT 714

then performs an FFT on the estimated channel impulse response matrix, $\hat{\underline{\underline{\mathcal{H}}}}$, to obtain the estimated channel frequency response matrix, $\hat{\underline{\underline{\mathbf{H}}}}$ (i.e., $\hat{\underline{\underline{\mathbf{H}}}} = \text{FFT}[\hat{\underline{\underline{\mathcal{H}}}}]$).

[1103] A unit 716 then performs the channel eigen-decomposition of $\hat{\underline{\mathbf{H}}}(k)$, for each frequency bin k, to obtain the corresponding matrix of left eigen-vectors, $\underline{\mathbf{U}}(k)$. Each column of $\underline{\underline{\mathbf{U}}}$, where $\underline{\underline{\mathbf{U}}} = [\underline{\mathbf{U}}(0) \ \ldots \ \underline{\mathbf{U}}(k) \ \ldots \ \underline{\mathbf{U}}(N_F - 1)]$, is a steering vector for a corresponding element of $\underline{\mathbf{r}}(n)$, and is used to orthogonalize the received symbol streams. An IFFT 718 then performs the inverse FFT of $\underline{\underline{\mathbf{U}}}$ to obtain a spatio-temporal pulse-shaping matrix, $\underline{\mathcal{U}}(n)$, for the receiver system.

[1104] A convolver 720 then conditions the received symbol vectors, $\underline{\mathbf{r}}(n)$, with the conjugate transpose of the spatio-temporal pulse-shaping matrix, $\underline{\mathcal{U}}^H(n)$, to obtain the recovered symbol vectors, $\hat{\underline{\mathbf{s}}}(n)$, which are estimates of the modulation symbol vectors, $\underline{\mathbf{s}}(n)$. In the time domain, the conditioning is a convolution operation, which may be expressed as:

$$\hat{\underline{\mathbf{s}}}(n) = \sum_{\ell} \underline{\mathcal{U}}^H(\ell)\,\underline{\mathbf{r}}(n-\ell) \ . \qquad \text{Eq (14)}$$

[1105] The pulse-shaping at the receiver may also be performed in the frequency domain, similar to that described above for the transmitter. In this case, the N_R sequences of received symbols for the N_R receive antennas, which make up the sequence of received symbol vectors, $\underline{\mathbf{r}}(n)$, are each broken up into subsequences of N_{SS} received symbols, and each subsequence is zero-padded to provide a corresponding vector of length N_O. The N_R sequences of $\underline{\mathbf{r}}(n)$ are thus processed to provide N_R sequences of vectors of length N_O, $\tilde{\underline{\mathbf{r}}}_\ell(n)$, where the subscript ℓ is the index for the vectors that correspond to the zero-padded subsequences. Each zero-padded subsequence is then transformed via an FFT, resulting in a sequence of frequency-domain vectors, $\underline{\mathbf{R}}_\ell(k)$, for different values of ℓ. Each vector $\underline{\mathbf{R}}_\ell(k)$, for a given ℓ, includes a set of frequency-domain vectors of length N_O (i.e., for $k = (0,\ 1,\ \ldots\ , N_O - 1)$).

[1106] The conjugate transpose of the spatio-temporal pulse-shaping matrix, $\underline{\mathcal{U}}^H(n)$, is also zero-padded and transformed via an FFT to obtain a frequency-domain

matrix, $\underline{\tilde{\mathbf{U}}}^{H}(k)$ for $k=(0,\ 1,\ \ldots\ ,N_O-1)$. The vector $\underline{\mathbf{R}}_{\ell}(k)$, for each value of ℓ, is then pre-multiplied with the conjugate transpose matrix $\underline{\tilde{\mathbf{U}}}^{H}(k)$ (where the pre-multiplication is performed for each value of k, i.e., for $k=(0,\ 1,\ \ldots\ ,N_O-1)$) to obtain a corresponding frequency-domain vector $\underline{\hat{\mathbf{S}}}_{\ell}(k)$. Each vector $\underline{\hat{\mathbf{S}}}_{\ell}(k)$, which includes a set of frequency-domain vectors of length N_O, can then be transformed via an inverse FFT to provide a corresponding set of time-domain subsequences of length N_O. The resulting subsequences are then reassembled according to the overlap-add method, or similar means, as is well-known in the art, to obtain sequences of recovered symbols, which corresponds to the set of recovered symbol vectors, $\underline{\hat{\mathbf{s}}}(n)$.

[1107] Thus recovered symbol vectors, $\underline{\hat{\mathbf{s}}}(n)$, may be characterized as a convolution in the time domain, as follows:

$$\underline{\hat{\mathbf{s}}}(n)=\sum_{\ell}\underline{\Gamma}(\ell)\,\underline{\mathbf{s}}(n-\ell)+\underline{\hat{\mathbf{z}}}(n)\ , \qquad \text{Eq (15)}$$

where $\underline{\Gamma}(\ell)$ is the inverse FFT of $\underline{\hat{\Lambda}}(k)=\underline{\Lambda}(k)\sqrt{\underline{\mathbf{W}}(k)}$; and

$\underline{\hat{\mathbf{z}}}(n)$ is the received noise as transformed by the receiver spatio-temporal pulse-shaping matrix, $\underline{\mathcal{U}}^{H}(\ell)$.

The matrix $\underline{\Gamma}(n)$ is a diagonal matrix of eigen-pulses derived from the set of matrices $\underline{\underline{\hat{\Lambda}}}$, where $\underline{\underline{\hat{\Lambda}}}=[\underline{\hat{\Lambda}}(0)\ \ldots\ \underline{\hat{\Lambda}}(k)\ \ldots\ \underline{\hat{\Lambda}}(N_F-1)]$. In particular, each diagonal element of $\underline{\Gamma}(n)$ corresponds to an eigen-pulse that is obtained as the IFFT of a set of singular values, $[\hat{\lambda}_{ii}(0)\ \ldots\ \hat{\lambda}_{ii}(k)\ \ldots\ \hat{\lambda}_{ii}(N_F-1)]^T$, for a corresponding element of $\underline{\underline{\hat{\Lambda}}}$,

[1108] The two forms for ordering the singular values, sorted and random-ordered, result in two different types of eigen-pulses. For the sorted form, the resulting eigen-pulse matrix, $\underline{\Gamma}_s(n)$, is a diagonal matrix of pulses that are sorted in descending order of energy content. The pulse corresponding to the first diagonal element of the eigen-pulse matrix, $\{\underline{\Gamma}_s(n)\}_{11}$, has the most energy, and the pulses corresponding to elements further down the diagonal have successively less energy. Furthermore, when the SNR is low enough that water-pouring results in some of the frequency bins having little or no energy, the energy is taken out of the smallest eigen-pulses first. Thus, at low SNRs,

one or more of the eigen-pulses may have little or no energy. This has the advantage that at low SNRs, the coding and modulation are simplified through the reduction in the number of orthogonal subchannels. However, in order to approach channel capacity, the coding and modulation are performed separately for each eigen-pulse.

[1109] The random-ordered form of the singular values in the frequency domain may be used to further simplify coding and modulation (i.e., to avoid the complexity of separate coding and modulation for each element of the eigen-pulse matrix). In the random-ordered form, for each frequency bin, the ordering of the singular values is random instead of being based on their magnitude or size. This random ordering can result in approximately equal energy in all of the eigen-pulses. When the SNR is low enough to result in frequency bins with little or no energy, these bins are spread approximately evenly among the eigenmodes so that the number of eigen-pulses with non-zero energy is the same independent of SNR. At high SNRs, the random-order form has the advantage that all of the eigen-pulses have approximately equal energy, in which case separate coding and modulation for different eigenmodes are not required.

[1110] If the response of the MIMO channel is frequency selective, then the elements in the diagonal matrices, $\underline{\Lambda}(k)$, are time-dispersive. However, because of the pre-processing at the transmitter to invert the channel, the resulting recovered symbol sequences, $\hat{\underline{s}}(n)$, have little intersymbol interference, if the channel inversion is effectively performed. In that case, additional equalization would not be required at the receiver to achieve high performance.

[1111] If the channel inversion is not effective (e.g., due to an inaccurate estimated channel frequency response matrix, $\hat{\underline{\underline{\mathbf{H}}}}$) then an equalizer may be used to equalize the recovered symbols, $\hat{\underline{s}}(n)$, prior to the demodulation and decoding. Various types of equalizer may be used to equalize the recovered symbol streams, including a minimum mean square error linear equalizer (MMSE-LE), a decision feedback equalizer (DFE), a maximum likelihood sequence estimator (MLSE), and so on.

[1112] Since the orthogonalization process at the transmitter and receiver results in decoupled (i.e., orthogonal) recovered symbol streams at the receiver, the complexity of the equalization required for the decoupled symbol streams is greatly reduced. In particular, the equalization may be achieved by parallel time-domain equalization of the independent symbol streams. The equalization may be performed as described in the

aforementioned U.S. Patent Application Serial No. 10/017,038, and in U.S. Patent Application Serial No. 09/993,087, entitled "Multiple-Access Multiple-Input Multiple-Output (MIMO) Communication System," filed November 6, 2001, which is assigned to the assignee of the present application and incorporated herein by reference.

[1113] For the embodiment in FIG. 7, the recovered symbol vectors, $\hat{\underline{s}}(n)$, are provided to RX data processor 162a. Within processor 162a, a symbol demapping element 732 demodulates each recovered symbol in $\hat{\underline{s}}(n)$ in accordance with a demodulation scheme that is complementary to the modulation scheme used for that symbol at the transmitter system. The demodulated data from symbol demapping element 732 is then de-interleaved by a deinterleaver 734. The deinterleaved data is further decoded by a decoder 736 to obtain the decoded bits, $\hat{d}_i$, which are estimates of the transmitted information bits, d_i. The deinterleaving and decoding are performed in a manner complementary to the interleaving and encoding, respectively, performed at the transmitter system. For example, a Turbo decoder or a Viterbi decoder may be used for decoder 736 if Turbo or convolutional coding, respectively, is performed at the transmitter system.

[1114] **FIG. 8** is a flow diagram of a process 800 that may be performed at the receiver unit to implement the various receive processing techniques described herein. Initially, an estimated channel response matrix for the MIMO channel is obtained (step 812). This matrix may be the estimated channel impulse response matrix, $\hat{\underline{\underline{\mathcal{H}}}}$, or the estimated channel frequency response matrix, $\hat{\underline{\underline{\mathbf{H}}}}$. The matrix $\hat{\underline{\underline{\mathcal{H}}}}$ or $\hat{\underline{\underline{\mathbf{H}}}}$ may be obtained, for example, based on pilot symbols transmitted over the MIMO channel. The estimated channel response matrix is then decomposed (e.g., using channel eigen-decomposition) to obtain a set of matrices of left eigen-vectors, $\underline{\underline{\mathbf{U}}}$ (step 814).

[1115] A pulse-shaping matrix $\underline{\underline{\mathcal{U}}}(n)$ is then derived based on the matrices of left eigen-vectors, $\underline{\underline{\mathbf{U}}}$ (step 816). The streams of received symbols are then conditioned (in either the time domain or frequency domain) based on the pulse-shaping matrix $\underline{\underline{\mathcal{U}}}(n)$ to provide the streams of recovered symbols (step 818). The recovered symbols are further processed in accordance with a particular receive processing scheme, which is

complementary to the transmit processing scheme used at the transmitter, to provide the decoded data (step 820).

[1116] Time-domain receive processing with channel eigenmode decomposition is described in further detail in the aforementioned U.S. Patent Application Serial No. 10/017,038.

[1117] The techniques for processing a data transmission at a transmitter and a receiver described herein may be implemented in various wireless communication systems, including but not limited to MIMO and CDMA systems. These techniques may also be used for the forward link and/or the reverse link.

[1118] The techniques described herein to process a data transmission at the transmitter and receiver may be implemented by various means. For example, these techniques may be implemented in hardware, software, or a combination thereof. For a hardware implementation, the elements used to perform various signal processing steps at the transmitter (e.g., to code and modulate the data, decompose the channel response matrix, derive the weights to invert the channel, derive the scaling values for power allocation, derive the transmitter pulse-shaping matrix, precondition the modulation symbols, and so on) or at the receiver (e.g., to decompose the channel response matrix, derive the receiver pulse-shaping matrix, condition the received symbols, demodulate and decode the recovered symbols, and so on) may be implemented within one or more application specific integrated circuits (ASICs), digital signal processors (DSPs), digital signal processing devices (DSPDs), programmable logic devices (PLDs), field programmable gate arrays (FPGAs), processors, controllers, micro-controllers, microprocessors, other electronic units designed to perform the functions described herein, or a combination thereof.

[1119] For a software implementation, some or all of the signal processing steps at each of the transmitter and receiver may be implemented with modules (e.g., procedures, functions, and so on) that perform the functions described herein. The software codes may be stored in a memory unit (e.g., memories 132 and 172 in FIG. 1) and executed by a processor (e.g., controllers 130 and 170). The memory unit may be implemented within the processor or external to the processor, in which case it can be communicatively coupled to the processor via various means as is known in the art.

[1120] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various

modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

[1121] **WHAT IS CLAIMED IS:**

CLAIMS

1. In a multiple-input multiple-output (MIMO) communication system, a method for processing data for transmission over a MIMO channel, comprising:

processing data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

deriving a pulse-shaping matrix based on an estimated response of the MIMO channel and in a manner to reduce intersymbol interference at a receiver; and

preconditioning the plurality of modulation symbol streams based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

2. The method of claim 1, further comprising:

deriving a plurality of weights based on an estimated channel response matrix for the MIMO channel, wherein the weights are used to invert a frequency response of the MIMO channel, and wherein the pulse-shaping matrix is further derived based on the weights.

3. The method of claim 2, further comprising:

decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors and a plurality of matrices of singular values, and

wherein the weights are derived based on the matrices of singular values and the pulse-shaping matrix is further derived based on the matrices of eigen-vectors.

4. The method of claim 2, wherein the estimated channel response matrix is descriptive of a plurality of eigenmodes of the MIMO channel.

5. The method of claim 4, wherein one set of weights is derived for each eigenmode used for data transmission and wherein the weights in each set are derived to invert the frequency response of the corresponding eigenmode.

6. The method of claim 4, further comprising:

deriving a plurality of scaling values based on the matrices of singular values, wherein the scaling values are used to adjust transmit powers for the eigenmodes of the MIMO channel, and wherein the pulse-shaping matrix is further derived based on the scaling values.

7. The method of claim 6, wherein the scaling values are derived based on water-pouring analysis.

8. The method of claim 3, wherein the estimated channel response matrix is provided in the frequency domain and is decomposed in the frequency domain.

9. The method of claim 3, wherein the estimated channel response matrix is decomposed using channel eigen-decomposition.

10. The method of claim 4, wherein eigenmodes associated with transmission capabilities below a particular threshold are not used for data transmission.

11. The method of claim 3, wherein the singular values in each matrix are sorted based on their magnitude.

12. The method of claim 4, wherein the singular values in each matrix are randomly ordered such that the eigenmodes of the MIMO channel are associated with approximately equal transmission capabilities.

13. The method of claim 1, wherein the pulse-shaping matrix comprises a plurality of sequences of time-domain values, and wherein the preconditioning is performed in the time domain by convolving the streams of modulation symbols with the pulse-shaping matrix.

14. The method of claim 1, wherein the pulse-shaping matrix comprises a plurality of sequences of frequency-domain values, and wherein the preconditioning is performed in the frequency domain by multiplying a plurality of streams of transformed modulation symbols with the pulse-shaping matrix.

15. The method of claim 1, wherein the pulse-shaping matrix is derived to maximize capacity by allocating more transmit power to eigenmodes of the MIMO channel having greater transmission capabilities.

16. The method of claim 1, wherein the pulse-shaping matrix is derived to provide approximately equal received signal-to-noise-and-interference ratios (SNRs) for the plurality of modulation symbol streams at the receiver.

17. The method of claim 1, wherein the particular processing scheme defines a separate coding and modulation scheme for each modulation symbol stream.

18. The method of claim 1, wherein the particular processing scheme defines a common coding and modulation scheme for all modulation symbol streams.

19. In a multiple-input multiple-output (MIMO) communication system, a method for processing data for transmission over a MIMO channel, comprising:

processing data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

obtaining an estimated channel response matrix for the MIMO channel;

decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors and a plurality of matrices of singular values;

deriving a plurality of weights based on the matrices of singular values, wherein the weights are used to invert the frequency response of the MIMO channel;

deriving a pulse-shaping matrix based on the matrices of eigen-vectors and the weights; and

preconditioning the plurality of streams of modulation symbols based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

20. The method of claim 19, further comprising:

deriving a plurality of scaling values based on the matrices of singular values, wherein the scaling values are used to adjust transmit powers for eigenmodes of the

MIMO channel, and wherein the pulse-shaping matrix is further derived based on the scaling values.

21. A memory communicatively coupled to a digital signal processing device (DSPD) capable of interpreting digital information to:

process data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

derive a pulse-shaping matrix based on an estimated response of the MIMO channel and in a manner to reduce intersymbol interference at a receiver; and

precondition the plurality of streams of modulation symbols based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

22. In a multiple-input multiple-output (MIMO) communication system, a method for processing a data transmission received via a MIMO channel, comprising:

obtaining an estimated channel response matrix for the MIMO channel;

decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

deriving a pulse-shaping matrix based on the matrices of eigen-vectors; and

conditioning a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver.

23. The method of claim 22, wherein the conditioning is performed in the time domain based on a time-domain pulse-shaping matrix.

24. The method of claim 22, wherein the conditioning is performed in the frequency domain and includes

transforming the plurality of received symbol streams to the frequency domain;

multiplying the transformed received symbol streams with a frequency-domain pulse-shaping matrix to provide a plurality of conditioned symbol streams; and

transforming the plurality of conditioned symbol streams to the time domain to provide the plurality of recovered symbol streams.

25. The method of claim 22, wherein the conditioning orthogonalizes a plurality of streams of modulation symbols transmitted over the MIMO channel.

26. The method of claim 22, further comprising:

demodulating the plurality of recovered symbol streams in accordance with one or more demodulation schemes to provide a plurality of demodulated data streams; and

decoding the plurality of demodulated data streams in accordance with one or more decoding schemes to provide decoded data.

27. The method of claim 22, further comprising:

deriving channel state information (CSI) comprised of the estimated channel response matrix for the MIMO channel; and

sending the CSI back to the transmitter.

28. In a multiple-input multiple-output (MIMO) communication system, a method for processing a data transmission received via a MIMO channel, comprising:

obtaining an estimated channel response matrix for the MIMO channel;

decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

deriving a pulse-shaping matrix based on the matrices of eigen-vectors;

conditioning a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver;

demodulating the plurality of recovered symbol streams in accordance with one or more demodulation schemes to provide a plurality of demodulated data streams; and

decoding the plurality of demodulated data streams in accordance with one or more decoding schemes to provide decoded data.

29. A memory communicatively coupled to a digital signal processing device (DSPD) capable of interpreting digital information to:

obtain an estimated channel response matrix for a MIMO channel used for a data transmission;

decompose the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

derive a pulse-shaping matrix based on the matrices of eigen-vectors; and

condition a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver.

30. A transmitter unit in a multiple-input multiple-output (MIMO) communication system, comprising:

a TX data processor operative to process data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols; and

a TX MIMO processor operative to derive a pulse-shaping matrix based on an estimated response of a MIMO channel and in a manner to reduce intersymbol interference at a receiver, and to precondition the plurality of modulation symbol streams based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

31. The transmitter unit of claim 30, wherein the TX MIMO processor is further operative to derive a plurality of weights based on an estimated channel response matrix for the MIMO channel, wherein the weights are used to invert the frequency response of the MIMO channel, and wherein the pulse-shaping matrix is derived based in part on the weights.

32. The transmitter unit of claim 31, wherein the TX MIMO processor is further operative to decompose the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors and a plurality of matrices of singular values, and

wherein the weights are derived based on the matrices of singular values and the pulse-shaping matrix is further derived based on the matrices of eigen-vectors.

33. The transmitter unit of claim 31, wherein the TX MIMO processor is further operative to derive a plurality of scaling values used to adjust transmit powers for the eigenmodes of the MIMO channel, and wherein the pulse-shaping matrix is further derived based on the scaling values.

34. The transmitter unit of claim 33, wherein the scaling values are derived based on water-pouring analysis on a plurality of matrices of singular values obtained from the estimated channel response matrix.

35. An apparatus in a multiple-input multiple-output (MIMO) communication system, comprising:

means for processing data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

means for deriving a pulse-shaping matrix based on an estimated response of a MIMO channel and in a manner to reduce intersymbol interference at a receiver; and

means for preconditioning the plurality of modulation symbol streams based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

36. A digital signal processor comprising:

means for processing data in accordance with a particular processing scheme to provide a plurality of streams of modulation symbols;

means for deriving a pulse-shaping matrix based on an estimated response of a multiple-input multiple-output (MIMO) channel and in a manner to reduce intersymbol interference at a receiver; and

means for preconditioning the plurality of modulation symbol streams based on the pulse-shaping matrix to provide a plurality of streams of preconditioned symbols for transmission over the MIMO channel.

37. A receiver unit in a multiple-input multiple-output (MIMO) communication system, comprising:

an RX MIMO processor operative to obtain an estimated channel response matrix for a MIMO channel used for a data transmission, decompose the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors, derive a pulse-shaping matrix based on the matrices of eigen-vectors, and condition a plurality of streams of received symbols based on the pulse-shaping matrix to obtain a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted over the MIMO channel, wherein the modulation symbols were preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at the receiver unit; and

an RX data processor operative to process the plurality of recovered symbol streams in accordance with a particular processing scheme to provide decoded data.

38. The receiver unit of claim 37, wherein the RX MIMO processor is operative to condition the plurality of streams of received symbols in the time domain based on a time-domain pulse-shaping matrix.

39. An apparatus in a multiple-input multiple-output (MIMO) communication system, comprising:

means for obtaining an estimated channel response matrix for a MIMO channel used for a data transmission;

means for decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

means for deriving a pulse-shaping matrix based on the matrices of eigen-vectors; and

means for conditioning a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver.

40. A digital signal processor comprising:

means for obtaining an estimated channel response matrix for a multiple-input multiple-output (MIMO) channel used for a data transmission;

means for decomposing the estimated channel response matrix to obtain a plurality of matrices of eigen-vectors;

means for deriving a pulse-shaping matrix based on the matrices of eigen-vectors; and

means for conditioning a plurality of streams of received symbols based on the pulse-shaping matrix to provide a plurality of streams of recovered symbols which are estimates of modulation symbols transmitted for the data transmission, wherein the modulation symbols are preconditioned at a transmitter, prior to transmission over the MIMO channel, in a manner to reduce intersymbol interference at a receiver.

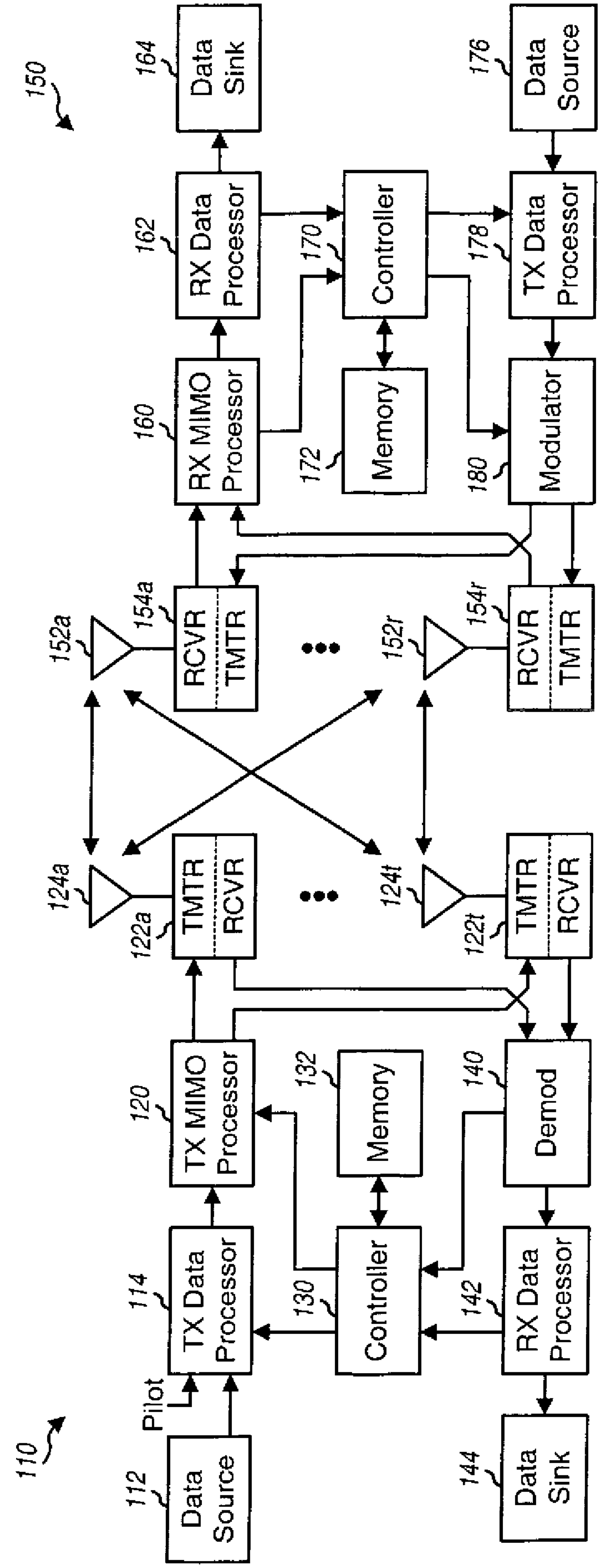

150
164
Data Sink
176
Data Source
162
RX Data Processor
170
Controller
178
TX Data Processor
160
RX MIMO Processor
172
Memory
180
Modulator
152a
154a
RCVR
TMTR
152r
154r
RCVR
TMTR
124a
122a
TMTR
RCVR
124t
122t
TMTR
RCVR
120
TX MIMO Processor
132
Memory
140
Demod
114
TX Data Processor
130
Controller
142
RX Data Processor
Pilot
110
112
Data Source
144
Data Sink

FIG. 1

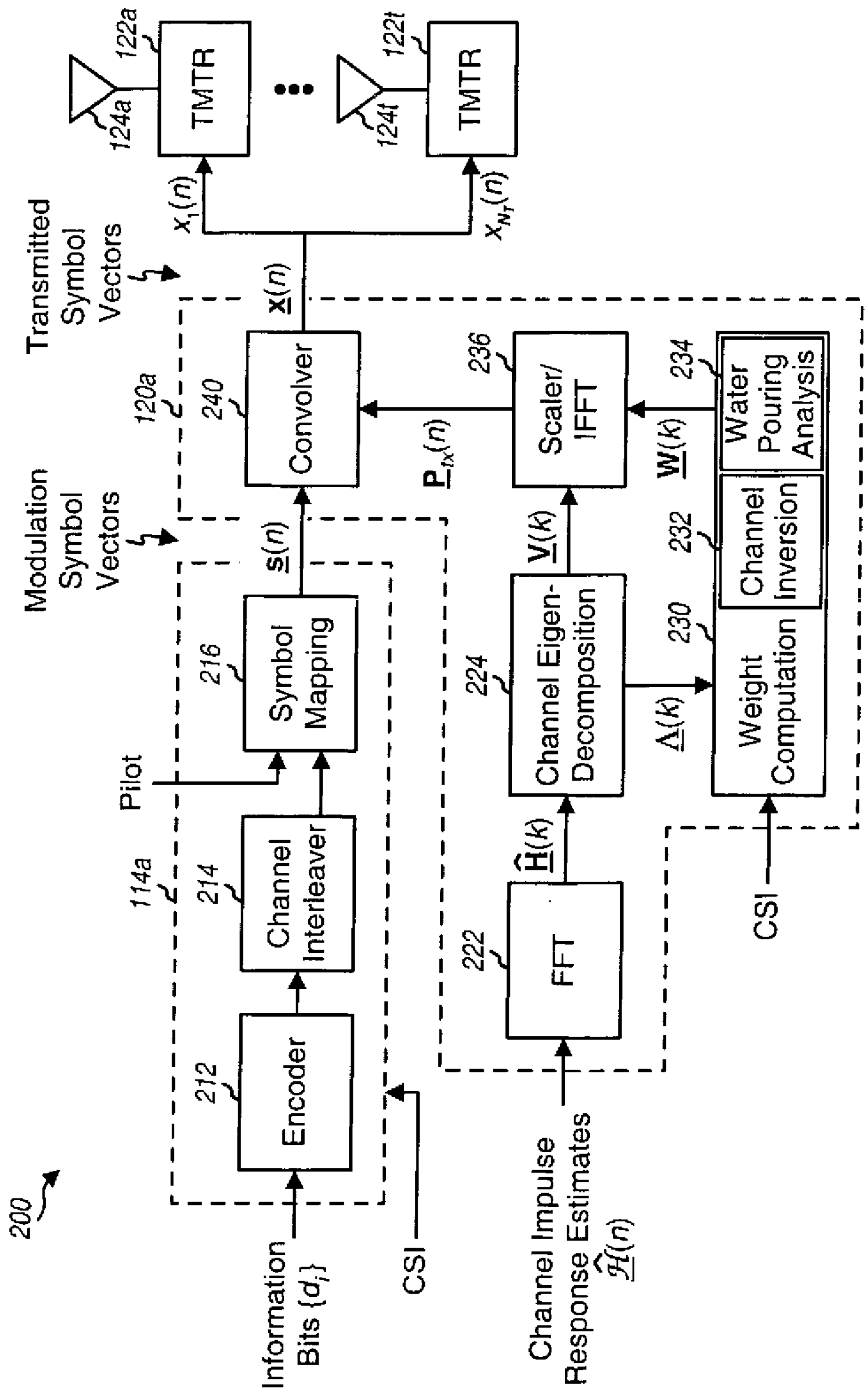
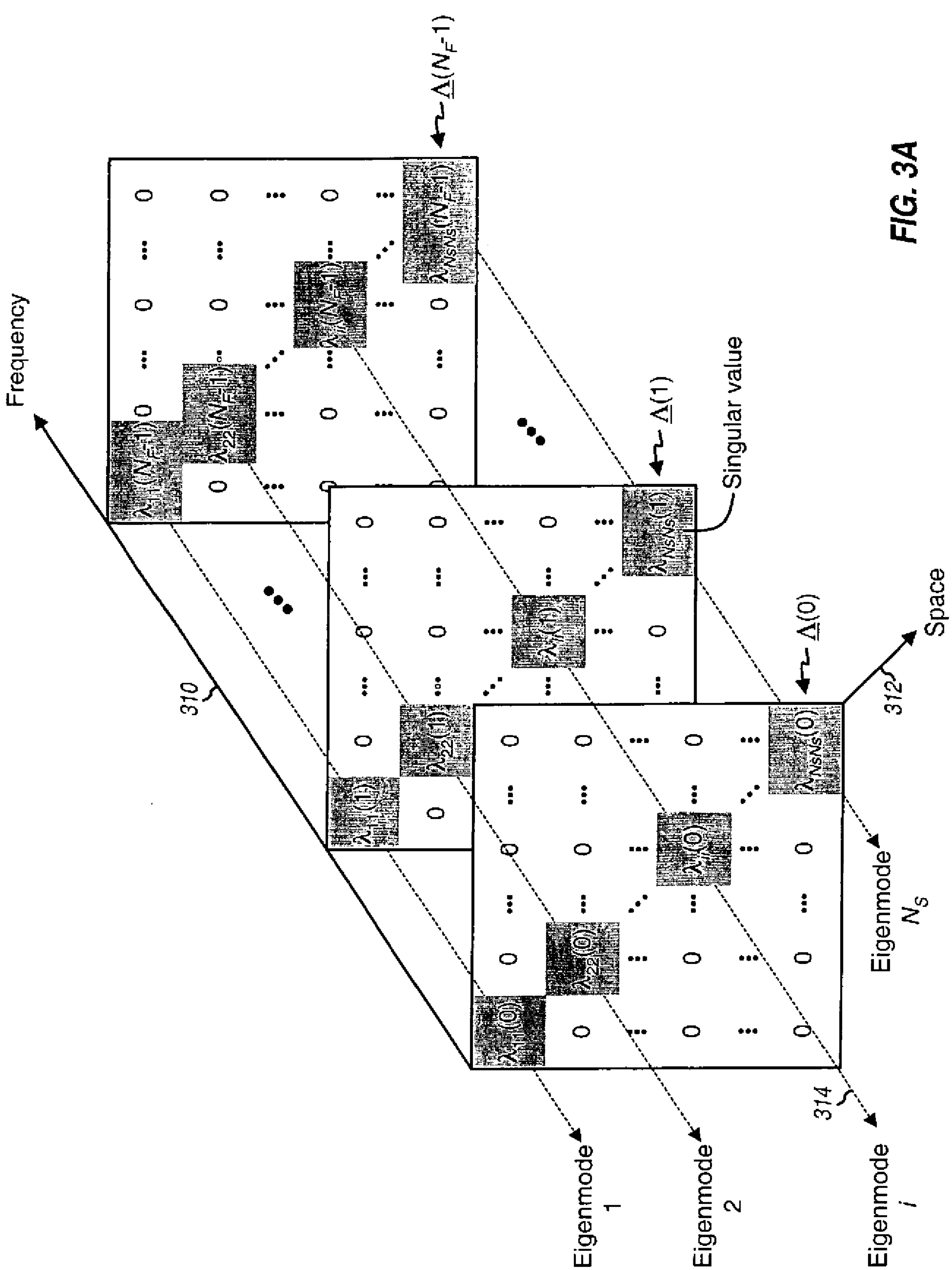

200
Information Bits $\{d_i\}$
CSI
114a
212
Encoder
214
Channel Interleaver
Pilot
216
Symbol Mapping
Modulation Symbol Vectors
$\underline{s}(n)$
120a
240
Convolver
Transmitted Symbol Vectors
$\underline{x}(n)$
$x_1(n)$
$x_{N_T}(n)$
122a
TMTR
124a
122t
TMTR
124t
Channel Impulse Response Estimates $\hat{\underline{\mathcal{H}}}(n)$
222
FFT
$\hat{\underline{H}}(k)$
224
Channel Eigen-Decomposition
$\underline{V}(k)$
$\underline{\Lambda}(k)$
236
Scaler/ IFFT
$\underline{P}_{tx}(n)$
230
Weight Computation
232
Channel Inversion
234
Water Pouring Analysis
$\underline{W}(k)$
CSI

FIG. 2



Λ(N_F-1)
Frequency
310
Singular value
Λ(1)
Λ(0)
Space
312
314
Eigenmode 1
Eigenmode 2
Eigenmode i
Eigenmode N_S

FIG. 3A

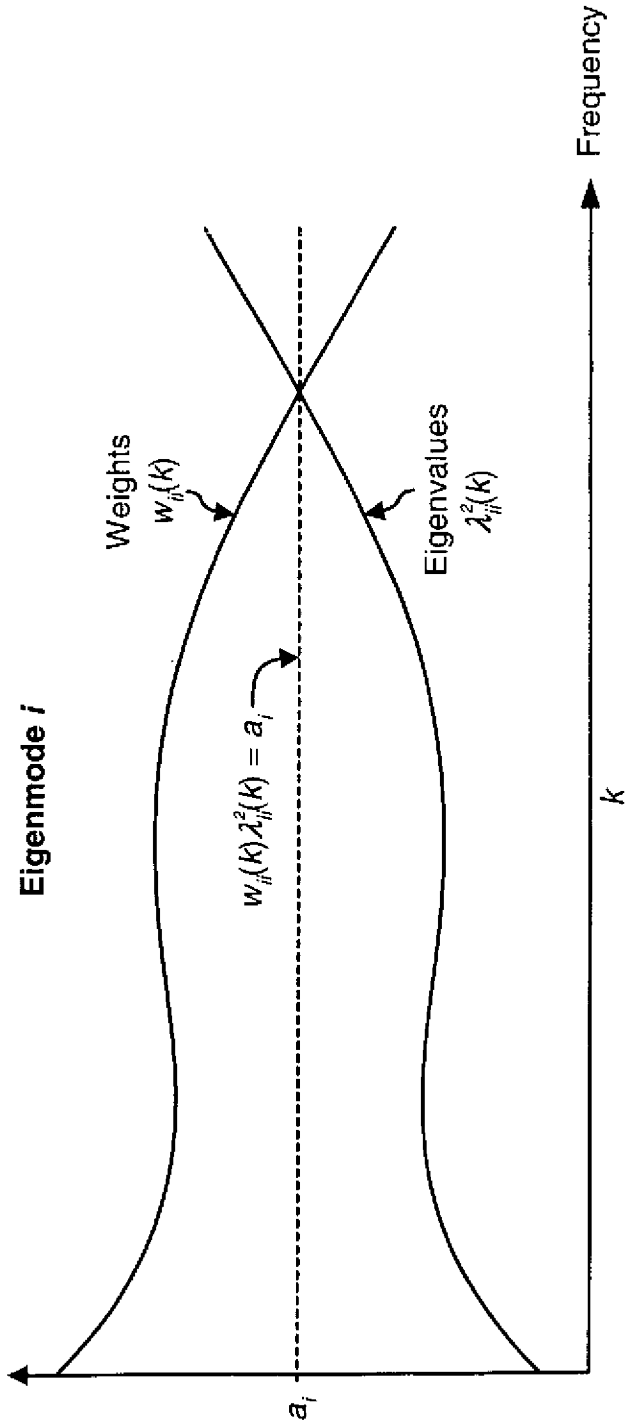
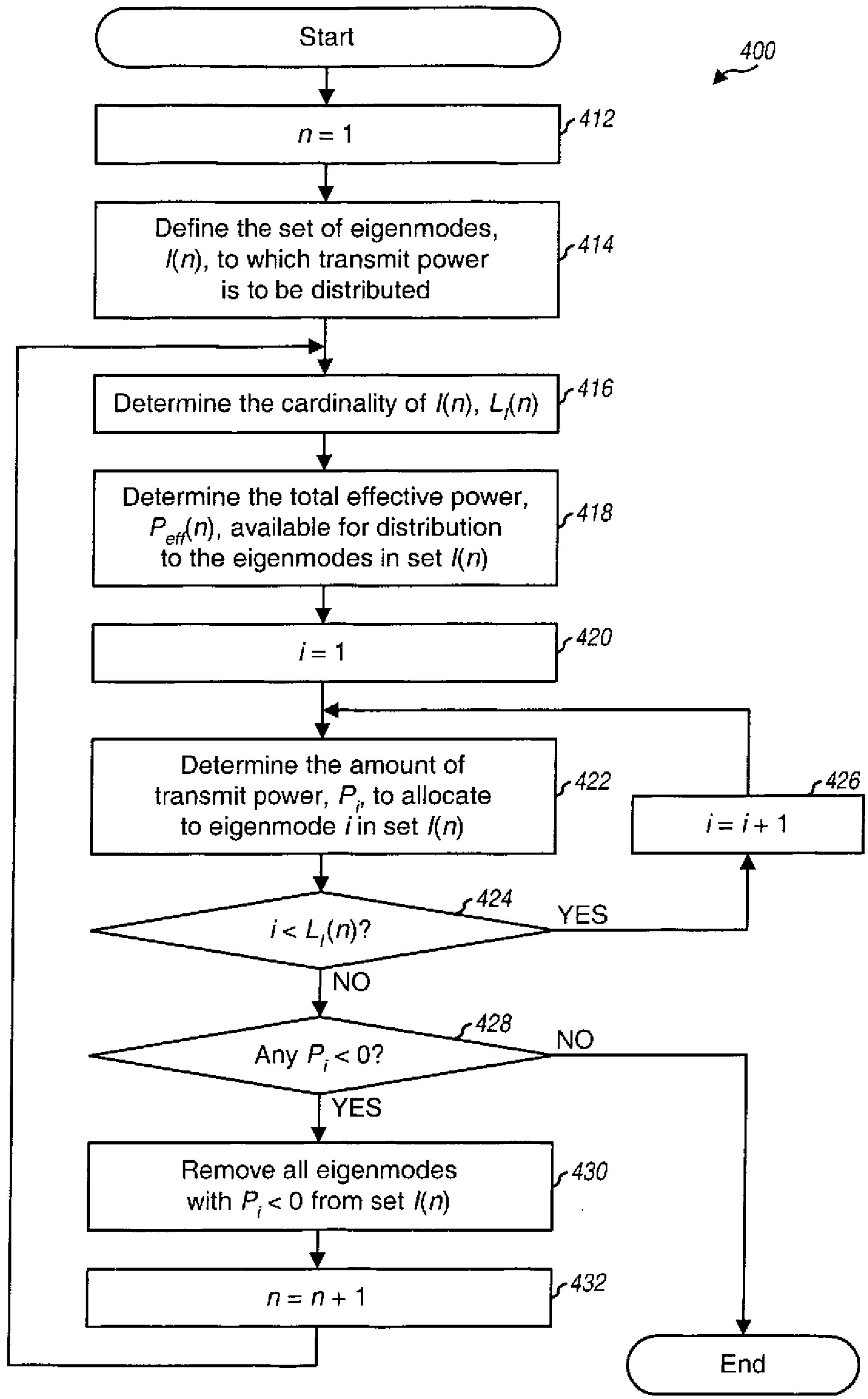
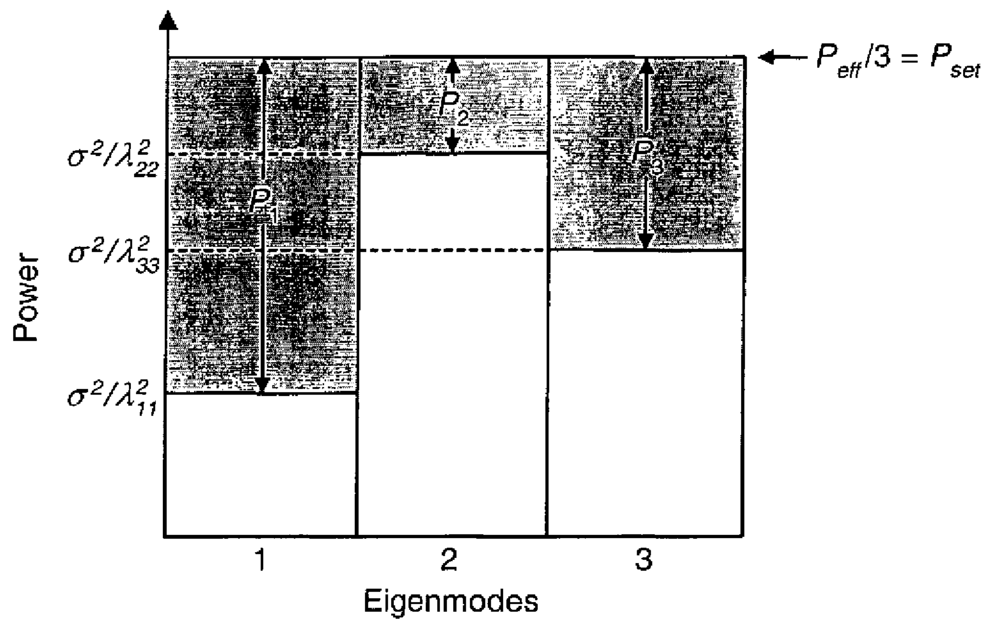

Eigenmode i
Weights
$w_{ii}(k)$
Eigenvalues
$\lambda_{ii}^2(k)$
$w_{ii}(k)\lambda_{ii}^2(k) = a_i$
a_i
k
Frequency

FIG. 3B



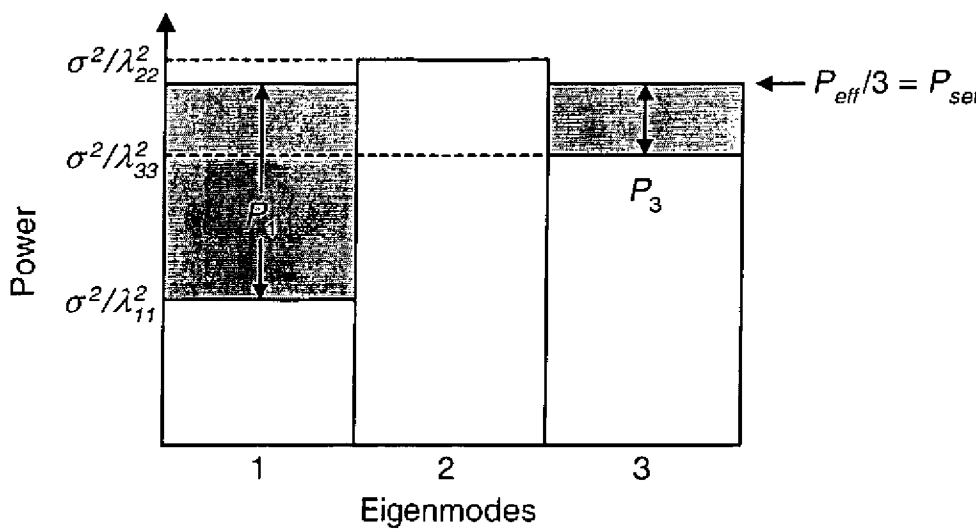
Start
400
n = 1
412
Define the set of eigenmodes, I(n), to which transmit power is to be distributed
414
Determine the cardinality of I(n), L_I(n)
416
Determine the total effective power, P_eff(n), available for distribution to the eigenmodes in set I(n)
418
i = 1
420
Determine the amount of transmit power, P_i, to allocate to eigenmode i in set I(n)
422
i = i + 1
426
i < L_I(n)?
424
YES
NO
Any P_i < 0?
428
NO
YES
Remove all eigenmodes with P_i < 0 from set I(n)
430
n = n + 1
432
End

FIG. 4



$P_{eff}/3 = P_{set}$
σ^2/λ_{22}^2
σ^2/λ_{33}^2
σ^2/λ_{11}^2
P_1
P_2
P_3
Power
1
2
3
Eigenmodes

FIG. 5A



σ^2/λ_{22}^2
$P_{eff}/3 = P_{set}$
σ^2/λ_{33}^2
σ^2/λ_{11}^2
P_1
P_3
Power
1
2
3
Eigenmodes

FIG. 5B

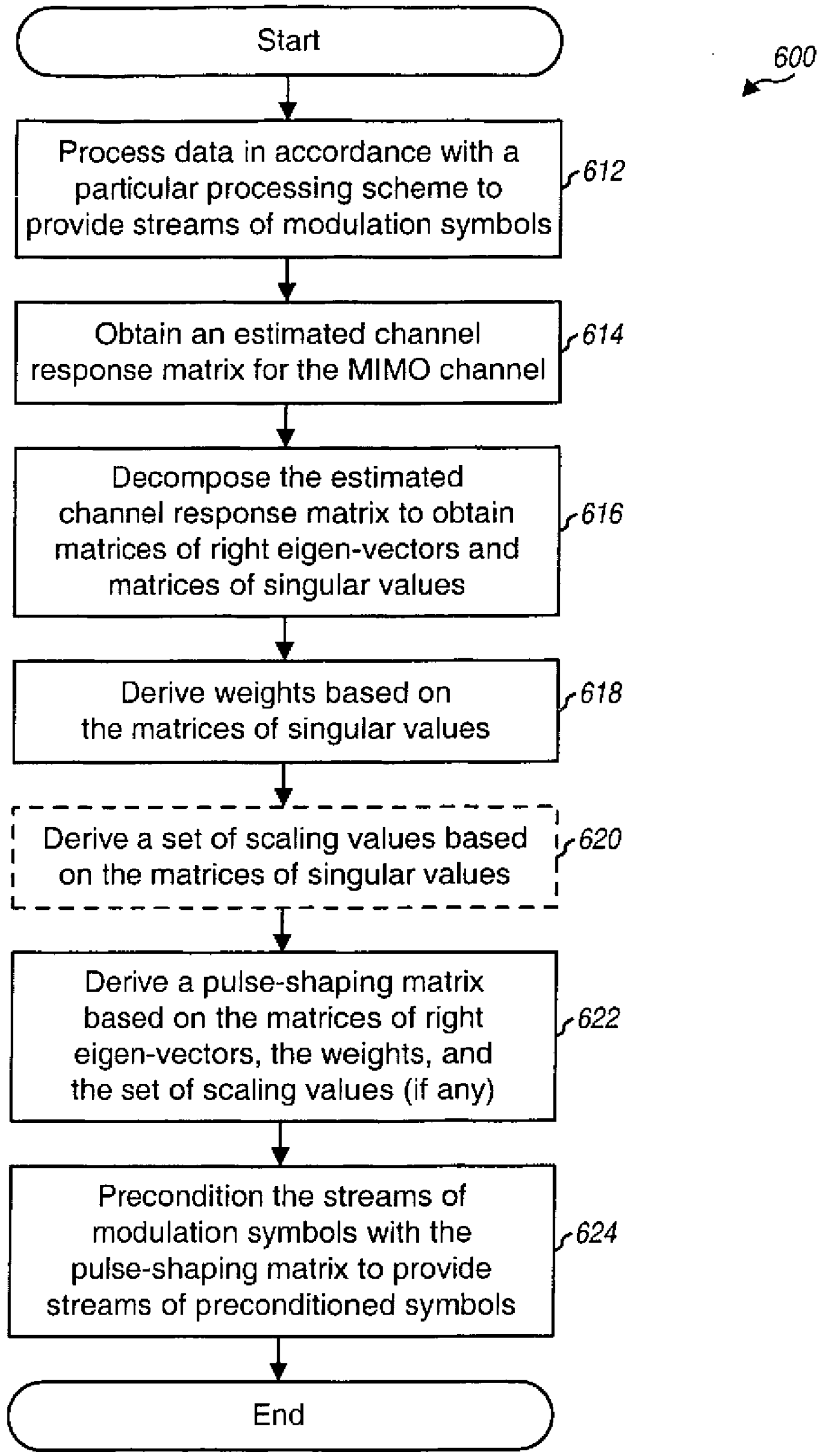

600
Start
Process data in accordance with a particular processing scheme to provide streams of modulation symbols
612
Obtain an estimated channel response matrix for the MIMO channel
614
Decompose the estimated channel response matrix to obtain matrices of right eigen-vectors and matrices of singular values
616
Derive weights based on the matrices of singular values
618
Derive a set of scaling values based on the matrices of singular values
620
Derive a pulse-shaping matrix based on the matrices of right eigen-vectors, the weights, and the set of scaling values (if any)
622
Precondition the streams of modulation symbols with the pulse-shaping matrix to provide streams of preconditioned symbols
624
End

FIG. 6

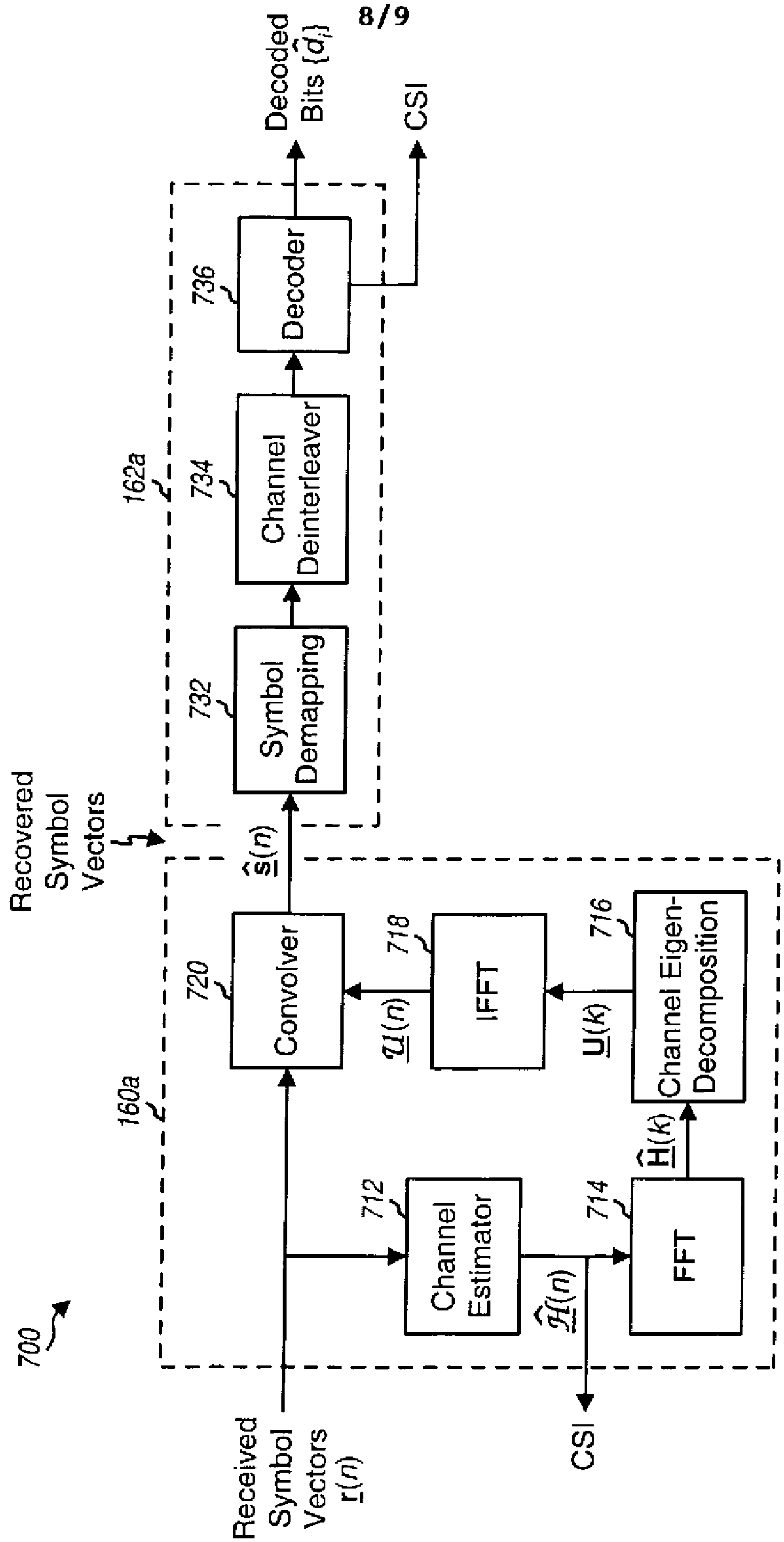

700
Received Symbol Vectors $\underline{r}(n)$
160a
Channel Estimator
712
$\hat{\mathcal{H}}(n)$
CSI
FFT
714
$\hat{\underline{H}}(k)$
Channel Eigen-Decomposition
716
$\underline{U}(k)$
IFFT
718
$\underline{\mathcal{U}}(n)$
Convolver
720
$\hat{\underline{s}}(n)$
Recovered Symbol Vectors
162a
Symbol Demapping
732
Channel Deinterleaver
734
Decoder
736
Decoded Bits $\{\hat{d}_i\}$
CSI

FIG. 7

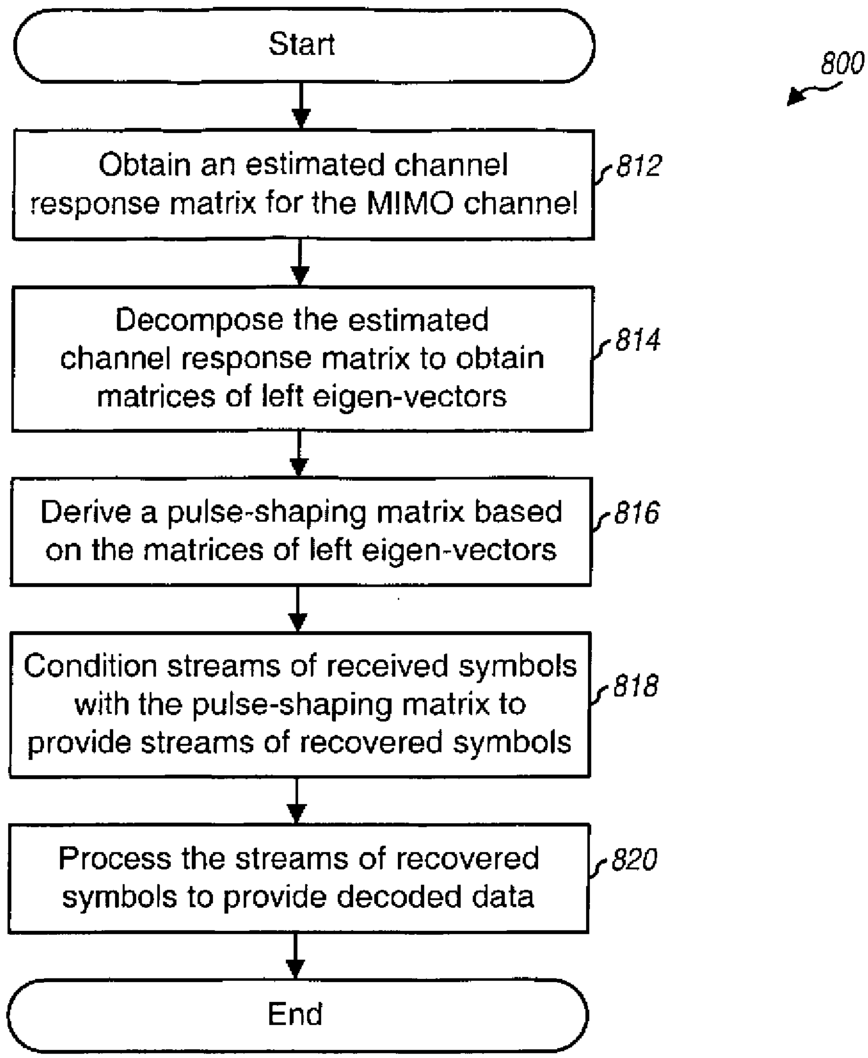

800
Start
Obtain an estimated channel response matrix for the MIMO channel
812
Decompose the estimated channel response matrix to obtain matrices of left eigen-vectors
814
Derive a pulse-shaping matrix based on the matrices of left eigen-vectors
816
Condition streams of received symbols with the pulse-shaping matrix to provide streams of recovered symbols
818
Process the streams of recovered symbols to provide decoded data
820
End

FIG. 8

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 03/19464

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L1/06 H04L25/03 H04L25/02

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, WPI Data, PAJ, INSPEC

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	SAMPATH H ET AL: "Joint transmit and receive optimization for high data rate wireless communication using multiple antennas" SIGNALS, SYSTEMS, AND COMPUTERS, 1999. CONFERENCE RECORD OF THE THIRTY-THIRD ASILOMAR CONFERENCE ON OCT. 24-27, 1999, PISCATAWAY, NJ, USA,IEEE, US, 24 October 1999 (1999-10-24), pages 215-219, XP010373976 ISBN: 0-7803-5700-0 abstract page 215, paragraph 2 page 215, paragraph 5 -page 216, paragraph 2 page 217, column 2 -page 218, paragraph 1	1,13-18, 21-30, 35-40
A	---	2-12,19, 20,31-34

-/--

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
24 September 2003	06/10/2003

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Reilly, D

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
X	WO 02 05506 A (QUALCOMM INC) 17 January 2002 (2002-01-17) abstract page 13, line 4 - line 11 page 14, line 24 -page 19, line 4	1,13-18, 21-30, 35-40
A		2-12,19, 20,31-34
X	JOONSUK KIM ET AL: "Transmission Optimization with a Space-Time Filter at Low SNR Wireless Environment" GLOBECOM'99, vol. 1B, 5 December 1999 (1999-12-05), pages 889-893, XP010373673 abstract page 890, column 1, last paragraph -page 892, column 1, paragraph 2	1,13-18, 21-30, 35-40
A		2-12,19, 20,31-34
X	BURR A G: "Adaptive space-time signal processing and coding" IEEE, vol. 2, 22 October 2000 (2000-10-22), pages 710-714, XP010531996 abstract page 710, column 2, paragraph 3 -page 712, column 1, paragraph 3	1,13-18, 21-30, 35-40
A		2-12,19, 20,31-34
A	EP 0 549 019 A (KONINKL PHILIPS ELECTRONICS NV) 30 June 1993 (1993-06-30) abstract page 2, line 25 - line 30	2-12,19, 20,31-34

INTERNATIONAL SEARCH REPORT

Information on patent family members

International Application No
PCT/US 03/19464

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
WO 0205506 A	17-01-2002	AU 7341101 A	21-01-2002
		EP 1299978 A2	09-04-2003
		WO 0205506 A2	17-01-2002
EP 0549019 A	30-06-1993	DE 69227356 D1	26-11-1998
		DE 69227356 T2	27-05-1999
		EP 0549019 A2	30-06-1993
		JP 5327558 A	10-12-1993
		KR 260809 B1	01-07-2000
		US 5353310 A	04-10-1994

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International Bureau

(43) International Publication Date
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WO 02/065664 A2

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(21) International Application Number: PCT/US02/05171

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(25) Filing Language: English

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(30) Priority Data:
09/788,259 15 February 2001 (15.02.2001) US

(71) Applicant: QUALCOMM INCORPORATED [US/US]; 5775 Morehouse Drive, San Diego, CA 92121-1714 (US).

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(74) Agents: WADSWORTH, Philip, R. et al.; Qualcomm Incorporated, 5775 Morehouse Drive, San Diego, CA 92121-1714 (US).

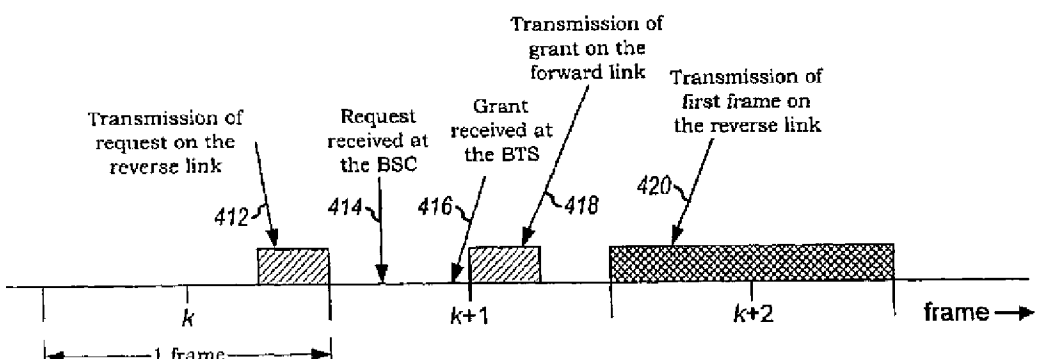
(81) Designated States (*national*): AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NO, NZ, OM, PH, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TN, TR, TT, TZ, UA, UG, UZ, VN, YU, ZA, ZM, ZW.

(84) Designated States (*regional*): ARIPO patent (GH, GM, KE, LS, MW, MZ, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian patent (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European patent (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE, TR), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Published:
— *without international search report and to be republished upon receipt of that report*

For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

(54) Title: REVERSE LINK CHANNEL ARCHITECTURE FOR A WIRELESS COMMUNICATION SYSTEM



(57) Abstract: A channel structure and mechanisms that support effective and efficient allocation and utilization of the reverse link resources. In one aspect, mechanisms are provided to quickly assign resources (e.g., a supplemental channel) as needed, and to quickly de-assign the resources when not needed or to maintain system stability. The reverse link resources may be quickly assigned and de-assigned via short messages (412, 418) exchanged on control channels on the forward and reverse links. In another aspect, mechanisms are provided to facilitate efficient and reliable data transmission. A reliable acknowledgment/negative acknowledgment scheme and an efficient retransmission scheme are provided. Mechanisms are also provided to control the transmit power and/or data rate of the remote terminals to achieve high performance and avoid instability.

WO 02/065664 A2

REVERSE LINK CHANNEL ARCHITECTURE FOR A WIRELESS COMMUNICATION SYSTEM

BACKGROUND

Field

[1001] The present invention relates generally to data communication, and more specifically to a novel and improved reverse link architecture for a wireless communication system.

Background

[1002] Wireless communication systems are widely deployed to provide various types of communication including voice and packet data services. These systems may be based on code division multiple access (CDMA), time division multiple access (TDMA), or some other modulation techniques. CDMA systems may provide certain advantages over other types of system, including increased system capacity.

[1003] In a wireless communication system, a user with a remote terminal (e.g., a cellular phone) communicates with another user through transmissions on the forward and reverse links via one or more base stations. The forward link (i.e., downlink) refers to transmission from the base station to the user terminal, and the reverse link (i.e., uplink) refers to transmission from the user terminal to the base station. The forward and reverse links are typically allocated different frequencies, a method called frequency division multiplexing (FDM).

[1004] The characteristics of packet data transmission on the forward and reverse links are typically very different. On the forward link, the base station usually knows whether or not it has data to transmit, the amount of data, and the identity of the recipient remote terminals. The base station may further be provided with the "efficiency" achieved by each recipient remote terminal, which may be quantified as the amount of transmit power needed per bit. Based on the known information, the base station may be able to efficiently schedule data

transmissions to the remote terminals at the times and data rates selected to achieve the desired performance.

[1005] On the reverse link, the base station typically does not know *a priori* which remote terminals have packet data to transmit, or how much. The base station is typically aware of each received remote terminal's efficiency, which may be quantified by the energy-per-bit-to-total-noise-plus-interface ratio, Ec/(No+Io), needed at the base station to correctly receive a data transmission. The base station may then allocate resources to the remote terminals whenever requested and as available.

[1006] Because of uncertainty in user demands, the usage on the reverse link may fluctuate widely. If many remote terminals transmit at the same time, high interference is generated at the base station. The transmit power from the remote terminals would need to be increased to maintain the target Ec/(No+Io), which would then result in higher levels of interference. If the transmit power is further increased in this manner, a "black out" may ultimately result and the transmissions from all or a large percentage of the remote terminals may not be properly received. This is due to the remote terminal not being able to transmit at sufficient power to close the link to the base station.

[1007] In a CDMA system, the channel loading on the reverse link is often characterized by what is referred to as the "rise-over-thermal". The rise-over-thermal is the ratio of the total received power at a base station receiver to the power of the thermal noise. Based on theoretical capacity calculations for a CDMA reverse link, there is a theoretical curve that shows the rise-over-thermal increasing with loading. The loading at which the rise-over-thermal is infinite is often referred to as the "pole". A loading that has a rise-over-thermal of 3 dB corresponds to a loading of about 50%, or about half of the number of users that can be supported when at the pole. As the number of users increases and as the data rates of the users increase, the loading becomes higher. Correspondingly, as the loading increases, the amount of power that a remote terminal must transmit increases. The rise-over-thermal and channel loading are described in further detail by A.J. Viterbi in "CDMA : Principles of Spread Spectrum Communication," Addison-Wesley Wireless Communications Series, May 1995, ISBN: 0201633744, which is incorporated herein by reference.

[1008] The Viterbi reference provides classical equations that show the relationship between the rise-over-thermal, the number of users, and the data rates of the users. The equations also show that there is greater capacity (in bits/second) if a few users transmit at a high rate than a larger number of users transmit at a higher rate. This is due to the interference between transmitting users.

[1009] In a typical CDMA system, many users' data rates are continuously changing. For example, in an IS-95 or cdma2000 system, a voice user typically transmits at one of four rates, corresponding to the voice activity at the remote terminal, as described in U.S Patent Nos. 5,657,420 and 5,778,338, both entitled "VARIABLE RATE VOCODER" and U.S Patent No. 5,742,734, entitled "ENCODING RATE SELECTION IN A VARIABLE RATE VOCODER". Similarly, many data users are continually varying their data rates. All this creates a considerable amount of variation in the amount of data being transmitted simultaneously, and hence a considerable variation in the rise-over-thermal.

[1010] As can be seen from the above, there is a need in the art for a reverse link channel structure capable of achieving high performance for packet data transmission, and which takes into consideration the data transmission characteristics of the reverse links.

SUMMARY

[1011] Aspects of the invention provide mechanisms that support effective and efficient allocation and utilization of the reverse link resources. In one aspect, mechanisms are provided to quickly assign resources (e.g., supplemental channels) as needed, and to quickly de-assign the resources when not needed or to maintain system stability. The reverse link resources may be quickly assigned and de-assigned via short messages exchanged on control channels on the forward and reverse links. In another aspect, mechanisms are provided to facilitate efficient and reliable data transmission. In particular, a reliable acknowledgment/negative acknowledgment scheme and an efficient retransmission scheme are provided. In yet another aspect,

mechanisms are provided to control the transmit power and/or data rate of the remote terminals to achieve high performance and avoid instability. Another aspect of the invention provides a channel structure capable of implementing the features described above. These and other aspects are described in further detail below.

[1012] The disclosed embodiments further provide methods, channel structures, and apparatus that implement various aspects, embodiments, and features of the invention, as described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

[1013] The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

[1014] FIG. 1 is a diagram of a wireless communication system that supports a number of users;

[1015] FIG. 2 is a simplified block diagram of an embodiment of a base station and a remote terminal;

[1016] FIGS. 3A and 3B are diagrams of a reverse and a forward channel structure, respectively;

[1017] FIG. 4 is a diagram illustrating a communication between the remote terminal and base station to assign a reverse link supplemental channel (R-SCH);

[1018] FIGS. 5A and 5B are diagrams illustrating a data transmission on the reverse link and an Ack/Nak message transmission for two different scenarios;

[1019] FIGS. 6A and 6B are diagrams illustrating an acknowledgment sequencing with short and long acknowledgment delays, respectively;

[1020] FIG. 7 is a flow diagram that illustrates a variable rate data transmission on the R-SCH with fast congestion control, in accordance with an embodiment of the invention; and

[1021] FIG. 8 is a diagram illustrating improvement that may be possible with fast control of the R-SCH.

DETAILED DESCRIPTION

[1022] FIG. 1 is a diagram of a wireless communication system 100 that supports a number of users and capable of implementing various aspects of the invention. System 100 provides communication for a number of cells, with each cell being serviced by a corresponding base station 104. The base stations are also commonly referred to as base transceiver systems (BTSs). Various remote terminals 106 are dispersed throughout the system. Each remote terminal 106 may communicate with one or more base stations 104 on the forward and reverse links at any particular moment, depending on whether or not the remote terminal is active and whether or not it is in soft handoff. The forward link refers to transmission from base station 104 to remote terminal 106, and the reverse link refers to transmission from remote terminal 106 to base station 104. As shown in FIG. 1, base station 104a communicates with remote terminals 106a, 106b, 106c, and 106d, and base station 104b communicates with remote terminals 106d, 106e, and 106f. Remote terminal 106d is in soft handoff and concurrently communicates with base stations 104a and 104b.

[1023] In system 100, a base station controller (BSC) 102 couples to base stations 104 and may further couple to a public switched telephone network (PSTN). The coupling to the PSTN is typically achieved via a mobile switching center (MSC), which is not shown in FIG. 1 for simplicity. The BSC may also couple into a packet network, which is typically achieved via a packet data serving node (PDSN) that is also not shown in FIG. 1. BSC 102 provides coordination and control for the base stations coupled to it. BSC 102 further controls the routing of telephone calls among remote terminals 106, and between remote terminals 106 and users coupled to the PSTN (e.g., conventional telephones) and to the packet network, via base stations 104.

[1024] System 100 may be designed to support one or more CDMA standards such as (1) the "TIA/EIA-95-B Mobile Station-Base Station Compatibility Standard for Dual-Mode Wideband Spread Spectrum Cellular System" (the IS-95 standard), (2) the "TIA/EIA-98-D Recommended Minimum Standard for Dual-Mode Wideband Spread Spectrum Cellular Mobile Station"

(the IS-98 standard), (3) the documents offered by a consortium named "3rd Generation Partnership Project" (3GPP) and embodied in a set of documents including Document Nos. 3G TS 25.211, 3G TS 25.212, 3G TS 25.213, and 3G TS 25.214 (the W-CDMA standard), (4) the documents offered by a consortium named "3rd Generation Partnership Project 2" (3GPP2) and embodied in a set of documents including Document Nos. C.S0002-A, C.S0005-A, C.S0010-A, C.S0011-A. C.S0024, and C.S0026 (the cdma2000 standard), and (5) some other standards. In the case of the 3GPP and 3GPP2 documents, these are converted by standards bodies worldwide (e.g., TIA, ETSI, ARIB, TTA, and CWTS) into regional standards and have been converted into international standards by the International Telecommunications Union (ITU). These standards are incorporated herein by reference.

[1025] FIG. 2 is a simplified block diagram of an embodiment of base station 104 and remote terminal 106, which are capable of implementing various aspects of the invention. For a particular communication, voice data, packet data, and/or messages may be exchanged between base station 104 and remote terminal 106. Various types of messages may be transmitted such as messages used to establish a communication session between the base station and remote terminal and messages used to control a data transmission (e.g., power control, data rate information, acknowledgment, and so on). Some of these message types are described in further detail below.

[1026] For the reverse link, at remote terminal 106, voice and/or packet data (e.g., from a data source 210) and messages (e.g., from a controller 230) are provided to a transmit (TX) data processor 212, which formats and encodes the data and messages with one or more coding schemes to generate coded data. Each coding scheme may include any combination of cyclic redundancy check (CRC), convolutional, Turbo, block, and other coding, or no coding at all. Typically, voice data, packet data, and messages are coded using different schemes, and different types of message may also be coded differently.

[1027] The coded data is then provided to a modulator (MOD) 214 and further processed (e.g., covered, spread with short PN sequences, and scrambled with a long PN sequence assigned to the user terminal). The modulated data is then provided to a transmitter unit (TMTR) 216 and

conditioned (e.g., converted to one or more analog signals, amplified, filtered, and quadrature modulated) to generate a reverse link signal. The reverse link signal is routed through a duplexer (D) 218 and transmitted via an antenna 220 to base station 104.

[1028] At base station 104, the reverse link signal is received by an antenna 250, routed through a duplexer 252, and provided to a receiver unit (RCVR) 254. Receiver unit 254 conditions (e.g., filters, amplifies, downconverts, and digitizes) the received signal and provides samples. A demodulator (DEMOD) 256 receives and processes (e.g., despreads, decovers, and pilot demodulates) the samples to provide recovered symbols. Demodulator 256 may implement a rake receiver that processes multiple instances of the received signal and generates combined symbols. A receive (RX) data processor 258 then decodes the symbols to recover the data and messages transmitted on the reverse link. The recovered voice/packet data is provided to a data sink 260 and the recovered messages may be provided to a controller 270. The processing by demodulator 256 and RX data processor 258 are complementary to that performed at remote terminal 106. Demodulator 256 and RX data processor 258 may further be operated to process multiple transmissions received via multiple channels, e.g., a reverse fundamental channel (R-FCH) and a reverse supplemental channel (R-SCH). Also, transmissions may be received simultaneously from multiple remote terminals, each of which may be transmitting on a reverse fundamental channel, a reverse supplemental channel, or both.

[1029] On the forward link, at base station 104, voice and/or packet data (e.g., from a data source 262) and messages (e.g., from controller 270) are processed (e.g., formatted and encoded) by a transmit (TX) data processor 264, further processed (e.g., covered and spread) by a modulator (MOD) 266, and conditioned (e.g., converted to analog signals, amplified, filtered, and quadrature modulated) by a transmitter unit (TMTR) 268 to generate a forward link signal. The forward link signal is routed through duplexer 252 and transmitted via antenna 250 to remote terminal 106.

[1030] At remote terminal 106, the forward link signal is received by antenna 220, routed through duplexer 218, and provided to a receiver unit 222.

Receiver unit 222 conditions (e.g., downconverts, filters, amplifies, quadrature demodulates, and digitizes) the received signal and provides samples. The samples are processed (e.g., despreaded, decovered, and pilot demodulated) by a demodulator 224 to provide symbols, and the symbols are further processed (e.g., decoded and checked) by a receive data processor 226 to recover the data and messages transmitted on the forward link. The recovered data is provided to a data sink 228, and the recovered messages may be provided to controller 230.

[1031] The reverse link has some characteristics that are very different from those of the forward link. In particular, the data transmission characteristics, soft handoff behaviors, and fading phenomenon are typically very different between the forward and reverse links.

[1032] As noted above, on the reverse link, the base station typically does not know *a priori* which remote terminals have packet data to transmit, or how much. Thus, the base station may allocate resources to the remote terminals whenever requested and as available. Because of uncertainty in user demands, the usage on the reverse link may fluctuate widely.

[1033] In accordance with aspects of the invention, mechanisms are provided to effectively and efficiently allocate and utilize the reverse link resources. In one aspect, mechanisms are provided to quickly assign resources as needed, and to quickly de-assign resources when not needed or to maintain system stability. The reverse link resources may be assigned via a supplemental channel that is used for packet data transmission. In another aspect, mechanisms are provided to facilitate efficient and reliable data transmission. In particular, a reliable acknowledgment scheme and an efficient retransmission scheme are provided. In yet another aspect, mechanisms are provided to control the transmit power of the remote terminals to achieve high performance and avoid instability. These and other aspects are described in further detail below.

[1034] FIG. 3A is a diagram of an embodiment of a reverse channel structure capable of implementing various aspects of the invention. In this embodiment, the reverse channel structure includes an access channel, an enhanced access channel, a pilot channel (R-PICH), a common control channel (R-CCCH), a

dedicated control channel (R-DCCH), a fundamental channel (R-FCH), supplemental channels (R-SCH), and a reverse rate indicator channel (R-RICH). Different, fewer, and/or additional channels may also be supported and are within the scope of the invention. These channels may be implemented similar to those defined by the cdma2000 standard. Features of some of these channels are described below.

[1035] For each communication (i.e., each call), a specific set of channels that may be used for the communication and their configurations are defined by one of a number of radio configurations (RC). Each RC defines a specific transmission format, which is characterized by various physical layer parameters such as, for example, the transmission rates, modulation characteristics, spreading rate, and so on. The radio configurations may be similar to those defined for the cdma2000 standard.

[1036] The reverse dedicated control channel (R-DCCH) is used to transmit user and signaling information (e.g., control information) to the base station during a communication. The R-DCCH may be implemented similar to the R-DCCH defined in the cdma2000 standard.

[1037] The reverse fundamental channel (R-FCH) is used to transmit user and signaling information (e.g., voice data) to the base station during a communication. The R-FCH may be implemented similar to the R-FCH defined in the cdma2000 standard.

[1038] The reverse supplemental channel (R-SCH) is used to transmit user information (e.g., packet data) to the base station during a communication. The R-SCH is supported by some radio configurations (e.g., RC3 through RC11), and is assigned to the remote terminals as needed and if available. In an embodiment, zero, one, or two supplemental channels (i.e., R-SCH1 and R-SCH2) may be assigned to the remote terminal at any given moment. In an embodiment, the R-SCH supports retransmission at the physical layer, and may utilize different coding schemes for the retransmission. For example, a retransmission may use a code rate of 1/2 for the original transmission. The same rate 1/2 code symbols may be repeated for the retransmission. In an alternative embodiment, the underlying code may be a rate 1/4 code. The original transmission may use 1/2 of the symbols and the retransmission may

use the other half of the symbols. If a third retransmission is done, it can repeat one of the group of symbols, part of each group, a subset of either group, and other possible combinations of symbols.

[1039] R-SCH2 may be used in conjunction with R-SCH1 (e.g., for RC11). In particular, R-SCH2 may be used to provide a different quality of service (QoS). Also, Type II and III hybrid ARQ schemes may be used in conjunction with the R-SCH. Hybrid ARQ schemes are generally described by S.B. Wicker in "Error Control System for Digital Communication and Storage," Prentice-Hall, 1995, Chapter 15, which is incorporated herein by reference. Hybrid ARQ schemes are also described in the cdma2000 standard.

[1040] The reverse rate indicator channel (R-RICH) is used by the remote terminal to provide information pertaining to the (packet) transmission rate on one or more reverse supplemental channels. Table 1 lists the fields for a specific format of the R-RICH. In an embodiment, for each data frame transmission on the R-SCH, the remote terminal sends a reverse rate indicator (RRI) symbol, which indicates the data rate for the data frame. The remote terminal also sends the sequence number of the data frame being transmitted, and whether the data frame is a first transmission or a retransmission. Different, fewer, and/or additional fields may also be used for the R-RICH and are within the scope of the invention. The information in Table 1 is sent by the remote terminal for each data frame transmitted on the supplemental channel (e.g., each 20 msec).

Table 1

Field	Length (bits)
RRI	3
SEQUENCE_NUM	2
RETRAN_NUM	2

[1041] If there are multiple reverse supplemental channels (e.g., R-SCH1 and R-SCH2), then there can be multiple R-RICH channels (e.g., R-RICH1 and R-RICH2), each with the RRI, SEQUENCE_NUM, and RETRAN_NUM fields. Alternatively, the fields for multiple reverse supplemental channels may be

combined into a single R-RICH channel. In a particular embodiment, the RRI field is not used, and fixed transmission rates are used or the base station performs blind rate determination in which the base determines the transmission rate from the data. Blind rate determination may be achieved in a manner described in U.S Patent No. 6,175,590, entitled "METHOD AND APPARATUS FOR DETERMINING THE RATE OF RECEIVED DATA IN A VARIABLE RATE COMMUNICATION SYSTEM," issued January 16, 2001, U.S Patent No. 5,751,725, entitled "METHOD AND APPARATUS FOR DETERMINING THE RATE OF RECEIVED DATA IN A VARIABLE RATE COMMUNICATION SYSTEM," issued May 12, 1998, both of which are assigned to the assignee of the present application and incorporated herein by reference.

[1042] FIG. 3B is a diagram of an embodiment of a forward channel structure capable of supporting various aspects of the invention. In this embodiment, the forward channel structure includes common channels, pilot channels, and dedicated channels. The common channels include a broadcast channel (F-BCCH), a quick paging channel (F-QPCH), a common control channel (F-CCCH), and a common power control channel (F-CPCCH). The pilot channels include a basic pilot channel and an auxiliary pilot channel. And the dedicated channels include a fundamental channel (F-FCH), a supplemental channel (F-SCH), a dedicated auxiliary channel (F-APICH), a dedicated control channel (F-DCCH), and a dedicated packet control channel (F-CPDCCH). Again, different, fewer, and/or additional channels may also be supported and are within the scope of the invention. These channels may be implemented similar to those defined by the cdma2000 standard. Features of some of these channels are described below.

[1043] The forward common power control channel (F-CPCCH) is used by the base station to transmit power control subchannels (e.g., one bit per subchannel) for power control of the R-PICH, R-FCH, R-DCCH, and R-SCH. In an embodiment, upon channel assignment, a remote terminal is assigned a reverse link power control subchannel from one of three sources - the F-DCCH, F-SCH, and F-CPCCH. The F-CPCCH may be assigned if the reverse link power control subchannel is not provided from either the F-DCCH or F-SCH.

[1044] In an embodiment, the available bits in the F-CPCCH may be used to form one or more power control subchannels, which may then be assigned for different uses. For example, a number of power control subchannels may be defined and used for power control of a number of reverse link channels. Power control for multiple channels based on multiple power control subchannels may be implemented as described in U.S. Patent No. 5,991,284, entitled "SUBCHANNEL POWER CONTROL," issued November 23, 1999, assigned to the assignee of the present application and incorporated herein by reference.

[1045] In one specific implementation, an 800 bps power control subchannel controls the power of the reverse pilot channel (R-PICH). All reverse traffic channels (e.g., the R-FCH, R-DCCH, and R-SCH) have their power levels related to the R-PICH by a known relationship, e.g., as described in C.S0002. The ratio between two channels is often referred to as the traffic-to-pilot ratio. The traffic-to-pilot ratio (i.e., the power level of the reverse traffic channel relative to the R-PICH) can be adjusted by messaging from the base station. However, this messaging is slow, so a 100 bits/second (bps) power control subchannel may be defined and used for power control of the R-SCH. In an embodiment, this R-SCH power control subchannel controls the R-SCH relative to the R-PICH. In another embodiment, the R-SCH power control subchannel controls the absolute transmission power of the R-SCH.

[1046] In an aspect of the invention, a "congestion" control subchannel may also be defined for control of the R-SCH, and this congestion control subchannel may be implemented based on the R-SCH power control subchannel or another subchannel.

[1047] Power control for the reverse link is described in further detail below.

[1048] The forward dedicated packet control channel (F-DPCCH) is used to transmit user and signaling information to a specific remote terminal during a communication. The F-DPCCH may be used to control a reverse link packet data transmission. In an embodiment, the F-DPCCH is encoded and interleaved to enhance reliability, and may be implemented similar to the F-DCCH defined by the cdma2000 standard.

[1049] Table 2 lists the fields for a specific format of the F-DPCCH. In an embodiment, the F-DPCCH has a frame size of 48 bits, of which 16 are used for CRC, 8 bits are used for the encoder tail, and 24 bits are available for data and messaging. In an embodiment, the default transmission rate for the F-DPCCH is 9600 bps, in which case a 48-bit frame can be transmitted in 5 msec time interval. In an embodiment, each transmission (i.e., each F-DPCCH frame) is covered with a public long code of the recipient remote terminal to which the frame is targeted. This avoids the need to use an explicit address (hence, the channel is referred to as a "dedicated" channel). However, the F-DPCCH is also "common" since a large number of remote terminals in dedicated channel mode may continually monitor the channel. If a message is directed to a particular remote terminal and is received correctly, then the CRC will check.

Table 2

Field	Number of Bits / Frame
Information	24
Frame Quality Indicator	16
Encoder Tail	8

[1050] The F-DPCCH may be used to transmit mini-messages, such as the ones defined by the cdma2000 standard. For example, the F-DPCCH may be used to transmit a *Reverse Supplemental Channel Assignment Mini Message* (RSCAMM) used to grant the F-SCH to the remote terminal.

[1051] The forward common packet Ack/Nak channel (F-CPANCH) is used by the base station to transmit (1) acknowledgments (Ack) and negative acknowledgments (Nak) for a reverse link packet data transmission and (2) other control information. In an embodiment, acknowledgments and negative acknowledgments are transmitted as n-bit Ack/Nak messages, with each message being associated with a corresponding data frame transmitted on the reverse link. In an embodiment, each Ack/Nak message may include 1, 2, 3, or 4 bits (or possible more bits), with the number of bits in the message being dependent on the number of reverse link channels in the service configuration.

The n-bit Ack/Nak message may be block coded to increase reliability or transmitted in the clear.

[1052] In an aspect, to improve reliability, the Ack/Nak message for a particular data frame is retransmitted in a subsequent frame (e.g., 20 msec later) to provide time diversity for the message. The time diversity provides additional reliability, or may allow for the reduction in power used to send the Ack/Nak message while maintaining the same reliability. The Ack/Nak message may use error correcting coding as is well known in the art. For the retransmission, the Ack/Nak message may repeat the exact same code word or may use incremental redundancy. Transmission and retransmission of the Ack/Nak is described in further detail below.

[1053] Several types of control are used on the forward link to control the reverse link. These include controls for supplemental channel request and grant, Ack/Nak for a reverse link data transmission, power control of the data transmission, and possibly others.

[1054] The reverse link may be operated to maintain the rise-over-thermal at the base station relatively constant as long as there is reverse link data to be transmitted. Transmission on the R-SCH may be allocated in various ways, two of which are described below:

- By infinite allocation. This method is used for real-time traffic that cannot tolerate much delay. The remote terminal is allowed to transmit immediately up to a certain allocated data rate.
- By scheduling. The remote terminal sends an estimate of its buffer size. The base station determines when the remote terminal is allowed to transmit. This method is used for available bit rate traffic. The goal of a scheduler is to limit the number of simultaneous transmissions so that the number of simultaneously transmitting remote terminals is limited, thus reducing the interference between remote terminals.

[1055] Since channel loading can change relatively dramatically, a fast control mechanism may be used to control the transmit power of the R-SCH (e.g., relative to the reverse pilot channel), as described below.

[1056] A communication between the remote terminal and base station to establish a connection may be achieved as follows. Initially, the remote terminal is in a dormant mode or is monitoring the common channels with the slotted timer active (i.e., the remote terminal is monitoring each slot). At a particular time, the remote terminal desires a data transmission and sends a short message to the base station requesting a reconnection of the link. In response, the base station may send a message specifying the parameters to be used for the communication and the configurations of various channels. This information may be sent via an *Extended Channel Assignment Message* (ECAM), a specially defined message, or some other message. This message may specify the following:

- The MAC_ID for each member of the remote terminal's Active Set or a subset of the Active Set. The MAC_ID is later used for addressing on the forward link.
- Whether the R-DCCH or R-FCH is used on the reverse link.
- For the F-CPANCH, the spreading (e.g., Walsh) codes and Active Set to be used. This may be achieved by (1) sending the spreading codes in the ECAM, or (2) transmitting the spreading codes in a broadcast message, which is received by the remote terminal. The spreading codes of neighbor cells may need to be included. If the same spreading codes can be used in neighboring cells, only a single spreading code may need to be sent.
- For the F-CPCCH, the Active Set, the channel identity, and the bit positions. In an embodiment, the MAC_ID may be hashed to the F-CPCCH bit positions to obviate the need to send the actual bit positions or subchannel ID to the remote terminal. This hashing is a pseudo-random method to map a MAC_ID to a subchannel on the F-CPCCH. Since different simultaneous remote terminals are assigned distinct MAC_IDs, the hashing can be such that these MAC_IDs also map to distinct F-CPCCH subchannels. For example, if there are K possible bit positions and N possible MAC_IDs, then $K = \_N \times ((40503 \times KEY) \bmod 2^{16}) / 2^{16}\_$, where KEY is the number that is fixed in this instance. There

are many other hash functions that can be used and discussions of such can be found in many textbooks dealing with computer algorithms.

[1057] In an embodiment, the message from the base station (e.g., the ECAM) is provided with a specific field, USE_OLD_SERV_CONFIG, used to indicate whether or not the parameters established in the last connection are to be used for the reconnection. This field can be used to obviate the need to send the *Service Connect Message* upon reconnection, which may reduce delay in re-establishing the connection.

[1058] Once the remote terminal has initialized the dedicated channel, it continues, for example, as described in the cdma2000 standard.

[1059] As noted above, better utilization of the reverse link resources may be achieved if the resources can be quickly allocated as needed and if available. In a wireless (and especially mobile) environment, the link conditions continually fluctuate, and long delay in allocating resources may result in inaccurate allocation and/or usage. Thus, in accordance with an aspect of the invention, mechanisms are provided to quickly assign and de-assign supplemental channels.

[1060] FIG. 4 is a diagram illustrating a communication between the remote terminal and base station to assign and de-assign a reverse link supplemental channel (R-SCH), in accordance with an embodiment of the invention. The R-SCH may be quickly assigned and de-assigned as needed. When the remote terminal has packet data to send that requires usage of the R-SCH, it requests the R-SCH by sending to the base station a *Supplemental Channel Request Mini Message* (SCRMM) (step 412). The SCRMM is a 5 msec message that may be sent on the R-DCCH or R-FCH. The base station receives the message and forwards it to the BSC (step 414). The request may or may not be granted. If the request is granted, the base station receives the grant (step 416) and transmits the R-SCH grant using a *Reverse Supplemental Channel Assignment Mini Message* (RSCAMM) (step 418). The RSCAMM is also a 5 msec message that may be sent on the F-FCH or F-DCCH (if allocated to the remote terminal) or on the F-DPCCH (otherwise). Once assigned, the remote terminal may thereafter transmit on the R-SCH (step 420).

[1061] Table 3 lists the fields for a specific format of the RSCAMM. In this embodiment, the RSCAMM includes 8 bits of layer 2 fields (i.e., the MSG_TYPE, ACK_SEQ, MSG_SEQ, and ACK_REQUIREMENT fields), 14 bits of layer 3 fields, and two reserved bits that are also used for padding as described in C.S0004 and C.S0005. The layer 3 (i.e., signaling layer) may be as defined in the cdma2000 standard.

Table 3

Field	Length (Bits)
MSG_TYPE	3
ACK_SEQUENCE	2
MSG_SEQUENCE	2
ACK_REQUIREMENT	1
REV_SCH_ID	1
REV_SCH_DURATION	4
REV_SCH_START_TIME	5
REV_SCH_NUM_BITS_IDX	4
RESERVED	2

[1062] When the remote terminal no longer has data to send on the R-SCH, it sends a *Resource Release Request Mini Message* (RRRMM) to the base station. If there is no additional signaling required between the remote terminal and base station, the base station responds with an *Extended Release Mini Message* (ERMM). The RRRMM and ERMM are also 5 msec messages that may be sent on the same channels used for sending the request and grant, respectively.

[1063] There are many scheduling algorithms that may be used to schedule the reverse link transmissions of remote terminals. These algorithms may tradeoff between rates, capacity, delay, error rates, and fairness (which gives all users some minimal level of services), to indicate some of the main criteria. In addition, the reverse link is subject to the power limitations of the remote terminal. In a single cell environment, the greatest capacity will exist when the smallest number of remote terminals is allowed to transmit with the highest rate that the remote terminal can support -- both in terms of capability and the ability

to provide sufficient power. However, in a multiple cell environment, it may be preferable for remote terminals near the boundary with another cell to transmit at a lower rate. This is because their transmissions cause interference into multiple cells -- not just a single cell. Another aspect that tends to maximize the reverse link capacity is to operate a high rise-over-thermal at the base station, which indicates high loading on the reverse link. It is for this reason that aspects of the invention use scheduling. The scheduling attempts to have a few number of remote terminals simultaneously transmit -- those that do transmit are allowed to transmit at the highest rates that they can support.

[1064] However, a high rise-over-thermal tends to result in less stability as the system is more sensitive to small changes in loading. It is for this reason that fast scheduling and control is important. Fast scheduling is important because the channel conditions change quickly. For instance, fading and shadowing processes may result in a signal that was weakly received at a base station suddenly becoming strong at the base station. For voice or certain data activity, the remote terminal autonomously changes the transmission rate While scheduling may be able to take some of this into account, scheduling may not be able to react sufficiently fast enough. For this reason, aspects of the invention provide fast power control techniques, which are described in further detail below.

[1065] An aspect of the invention provides a reliable acknowledgment/negative acknowledgment scheme to facilitate efficient and reliable data transmission. As described above, acknowledgments (Ack) and negative acknowledgments (Nak) are sent by the base station for data transmission on the R-SCH. The Ack/Nak can be sent using the F-CPANCH.

[1066] Table 4 shows a specific format for an Ack/Nak message. In this specific embodiment, the Ack/Nak message includes 4 bits that are assigned to four reverse link channels - the R-FCH, R-DCCH, R-SCH1, and R-SCH2. In an embodiment, an acknowledgment is represented by a bit value of zero ("0") and a negative acknowledgment is represented by a bit value of one ("1"). Other Ack/Nak message formats may also be used and are within the scope of the invention.

Table 4

Description	All Channels Used Number_Type (binary)	R-FCH, R-DCCH, and R-SCH1 Used Number_Type (binary)	R-FCH and R-DCCH Used Number_Type (binary)
ACK_R-FCH	xxx0	xxx0	xx00
NAK_R-FCH	xxx1	xxx1	xx11
ACK_R-DCCH	xx0x	xx0x	-
NAK_R-DCCH	xx1x	xx1x	-
ACK_R-SCH1	x0xx	00xx	00xx
NAK_R-SCH1	x1xx	11xx	11xx
ACK_R-SCH2	0xxx	-	-
NAK_R-SCH2	1xxx	-	-

[1067] In an embodiment, the Ack/Nak message is sent block coded but a CRC is not used to check for errors. This keeps the Ack/Nak message short and further allows the message to be sent with a small amount of energy. However, no coding may also be used for the Ack/Nak message, or a CRC may be attached to the message, and these variations are within the scope of the invention. In an embodiment, the base station sends an Ack/Nak message corresponding to each frame in which the remote terminal has been given permission to transmit on the R-SCH, and does not send Ack/Nak messages during frames that the remote terminal is not given permission to transmit.

[1068] During a packet data transmission, the remote terminal monitors the F-CPANCH for Ack/Nak messages that indicate the results of the transmission. The Ack/Nak messages may be transmitted from any number of base stations in the remote terminal's Active Set (e.g., from one or all base stations in the Active Set). The remote terminal can perform different actions depending on the received Ack/Nak messages. Some of these actions are described below.

[1069] If an Ack is received by the remote terminal, the data frame corresponding to the Ack may be removed from the remote terminal's physical layer transmit buffer (e.g., data source 210 in FIG. 2) since the data frame was correctly received by the base station.

[1070] If a Nak is received by the remote terminal, the data frame corresponding to the Nak may be retransmitted by the remote terminal if it is still in the physical layer transmit buffer. In an embodiment, there is a one-to-one correspondence between a forward link Ack/Nak message and a transmitted reverse link data frame. The remote terminal is thus able to identify the sequence number of the data frame not received correctly by the base station (i.e., the erased frame) based on the frame in which the Nak was received. If this data frame has not been discarded by the remote terminal, it may be retransmitted at the next available time interval, which is typically the next frame.

[1071] If neither an Ack nor a Nak was received, there are several next possible actions for the remote terminal. In one possible action, the data frame is maintained in the physical layer transmit buffer and retransmitted. If the retransmitted data frame is then correctly received at the base station, then the base station transmits an Ack. Upon correct receipt of this Ack, the remote terminal discards the data frame. This would be the best approach if the base station did not receive the reverse link transmission.

[1072] Another possible action is for the remote terminal to discard the data frame if neither an Ack nor a Nak was received. This would be the best alternative if the base station had received the frame but the Ack transmission was not received by the remote terminal. However, the remote terminal does not know the scenario that occurred and a policy needs to be chosen. One policy would be to ascertain the likelihood of the two events happening and performing the action that maximizes the system throughput.

[1073] In an embodiment, each Ack/Nak message is retransmitted a particular time later (e.g., at the next frame) to improve reliability of the Ack/Nak. Thus, if neither an Ack nor a Nak was received, the remote terminal combines the retransmitted Ack/Nak with the original Ack/Nak. Then, the remote terminal can proceed as described above. And if the combined Ack/Nak still does not result in a valid Ack or Nak, the remote terminal may discard the data frame and continue to transmit the next data frame in the sequence. The second transmission of the Ack/Nak may be at the same or lower power level relative to that of the first transmission.

[1074] If the base station did not actually receive the data frame after retransmissions, then a higher signaling layer at the base station may generate a message (e.g., an RLP NAK), which may result in the retransmission of the entire sequence of data frames that includes the erased frame.

[1075] FIG. 5A is a diagram illustrating a data transmission on the reverse link (e.g., the R-SCH) and an Ack/Nak transmission on the forward link. The remote terminal initially transmits a data frame, in frame *k*, on the reverse link (step 512). The base station receives and processes the data frame, and provides the demodulated frame to the BSC (step 514). If the remote terminal is in soft handoff, the BSC may also receive demodulated frames for the remote terminal from other base stations.

[1076] Based on the received demodulated frames, the BSC generates an Ack or a Nak for the data frame. The BSC then sends the Ack/Nak to the base station(s) (step 516), which then transmit the Ack/Nak to the remote terminal during frame k+1 (step 518). The Ack/Nak may be transmitted from one base station (e.g., the best base station) or from a number base stations in the remote terminal's Active Set. The remote terminal receives the Ack/Nak during frame k+1. If a Nak is received, the remote terminal retransmits the erased frame at the next available transmission time, which in this example is frame k+2 (step 520). Otherwise, the remote terminal transmits the next data frame in the sequence.

[1077] FIG. 5B is a diagram illustrating a data transmission on the reverse link and a second transmission of the Ack/Nak message. The remote terminal initially transmits a data frame, in frame *k*, on the reverse link (step 532). The base station receives and processes the data frame, and provides the demodulated frame to the BSC (step 534). Again, for soft handoff, the BSC may receive other demodulated frames for the remote terminal from other base stations.

[1078] Based on the received demodulated frames, the BSC generates an Ack or a Nak for the frame. The BSC then sends the Ack/Nak to the base station(s) (step 536), which then transmit the Ack/Nak to the remote terminal during frame k+1 (step 538). In this example, the remote terminal does not receive the Ack/Nak transmitted during frame k+1. However, the Ack/Nak for

the data frame transmitted in frame k is transmitted a second time during frame k+2, and is received by the remote terminal (step 540). If a Nak is received, the remote terminal retransmits the erased frame at the next available transmission time, which in this example is frame k+3 (step 542). Otherwise, the remote terminal transmits the next data frame in the sequence. As shown in FIG. 5B, the second transmission of the Ack/Nak improves the reliability of the feedback, and can result in improved performance for the reverse link.

[1079] In an alternative embodiment, the data frames are not sent back to the BSC from the base station, and the Ack/Nak is generated from the base station.

[1080] FIG. 6A is a diagram illustrating an acknowledgment sequencing with short acknowledgment delay. The remote terminal initially transmits a data frame with a sequence number of zero, in frame k, on the reverse link (step 612). For this example, the data frame is received in error at the base station, which then sends a Nak during frame k+1 (step 614). The remote terminal also monitors the F-CPANCH for an Ack/Nak message for each data frame transmitted on the reverse link. The remote terminal continues to transmit a data frame with a sequence number of one in frame k+1 (step 616).

[1081] Upon receiving the Nak in frame k+1, the remote terminal retransmits the erased frame with the sequence number of zero, in frame k+2 (step 618). The data frame transmitted in frame k+1 was received correctly, as indicated by an Ack received during frame k+2, and the remote terminal transmits a data frame with a sequence number of two in frame k+3 (step 620). Similarly, the data frame transmitted in frame k+2 was received correctly, as indicated by an Ack received during frame k+3, and the remote terminal transmits a data frame with a sequence number of three in frame k+4 (step 622). In frame k+5, the remote terminal transmits a data frame with a sequence number of zero for a new packet (step 624).

[1082] FIG. 6B is a diagram illustrating an acknowledgment sequencing with long acknowledgment delay such as when the remote terminal demodulates the Ack/Nak transmission based upon the retransmission of the Ack/Nak as described above. The remote terminal initially transmits a data frame with a sequence number of zero, in frame k, on the reverse link (step 632). The data

frame is received in error at the base station, which then sends a Nak (step 634). For this example, because of the longer processing delay, the Nak for frame *k* is transmitted during frame *k*+2. The remote terminal continues to transmit a data frame with a sequence number of one in frame *k*+1 (step 636) and a data frame with a sequence number of two in frame *k*+2 (step 638).

[1083] For this example, the remote terminal receives the Nak in frame *k*+2, but is not able to retransmit the erased frame at the next transmission interval. Instead, the remote terminal transmits a data frame with a sequence number of three in frame *k*+3 (step 640). At frame *k*+4, the remote terminal retransmits the erased frame with the sequence number of zero (step 642) since this frame is still in the physical layer buffer. Alternatively, the retransmission may be in frame *k*+3. And since the data frame transmitted in frame *k*+1 was received correctly, as indicated by an Ack received during frame *k*+3, and the remote terminal transmits a data frame with a sequence number of zero for a new packet (step 644).

[1084] As shown in FIG. 6B, the erased frame may be retransmitted at any time as long as it is still available in the buffer and there is no ambiguity as to which higher layer packet the data frame belongs to. The longer delay for the retransmission may be due to any number of reasons such as (1) longer delay to process and transmit the Nak, (2) non-detection of the first transmission of the Nak, (3) longer delay to retransmit the erased frame, and others.

[1085] An efficient and reliable Ack/Nak scheme can improve the utilization of the reverse link. A reliable Ack/Nak scheme may also allow data frames to be transmitted at lower transmit power. For example, without retransmission, a data frame needs to be transmitted at a higher power level (P_1) required to achieve one percent frame error rate (1% FER). If retransmission is used and is reliable, a data frame may be transmitted at a lower power level (P_2) required to achieve 10% FER. The 10% erased frames may be retransmitted to achieve an overall 1% FER for the transmission. Typically, $1.1 \cdot P_2 < P_1$, and less transmit power is used for a transmission using the retransmission scheme. Moreover, retransmission provides time diversity, which may improve performance. The retransmitted frame may also be combined with the first transmission of the frame at the base station, and the combined power from the two transmissions

may also improve performance. The recombining may allow an erased frame to be retransmitted at a lower power level.

[1086] An aspect of the invention provides various power control schemes for the reverse link. In an embodiment, reverse link power control is supported for the R-FCH, R-SCH, and R-DCCH. This can be achieved via a (e.g., 800 bps) power control channel, which may be partitioned into a number of power control subchannels. For example, a 100 bps power control subchannel may be defined and used for the R-SCH. If the remote terminal has not been allocated a F-FCH or F-DCCH, then the F-CPCCH may be used to send power control bits to the remote terminal.

[1087] In one implementation, the (e.g., 800 bps) power control channel is used to adjust the transmit power of the reverse link pilot. The transmit power of the other channels (e.g., the R-FCH) is set relative to that of the pilot (i.e., by a particular delta). Thus, the transmit power for all reverse link channels may be adjusted along with the pilot. The delta for each non-pilot channel may be adjusted by signaling. This implementation does not provide flexibility to quickly adjust the transmit power of different channels.

[1088] In one embodiment, the forward common power control channel (F-CPCCH) may be used to form one or more power control subchannels that may then be used for various purposes. Each power control subchannel may be defined using a number of available bits in the F-CPCCH (e.g., the m^{th} bit in each frame). For example, some of the available bits in the F-CPCCH may be allocated for a 100 bps power control subchannel for the R-SCH. This R-SCH power control subchannel may be assigned to the remote terminal during channel assignment. The R-SCH power control subchannel may then be used to (more quickly) adjust the transmit power of the designated R-SCH, e.g., relative to that of the pilot channel. For a remote terminal in soft handoff, the R-SCH power control may be based on the OR-of-the-downs rule, which decreases the transmit power if any base station in the remote terminal's Active Set directs a decrease. Since the power control is maintained at the base station, this permits the base station to adjust the transmitted power with minimal amount of delay and thus adjust the loading on the channel.

[1089] The R-SCH power control subchannel may be used in various manners to control the transmission on the R-SCH. In an embodiment, the R-SCH power control subchannel may be used to direct the remote terminal to adjust the transmit power on the R-SCH by a particular amount (e.g., 1 dB, 2 dB, or some other value). In another embodiment, the subchannel may be used to direct the remote terminal to reduce or increase transmit power by a large step (e.g., 3 dB, or possibly more). In both embodiments, the adjustment in transmit power may be relative to the pilot transmit power. In another embodiment, the subchannel may be directed to adjust the data rate allocated to the remote terminal (e.g., to the next higher or lower rate). In yet another embodiment, the subchannel may be used to direct the remote terminal to temporarily stop transmission. And in yet another embodiment, the remote terminal may apply different processing (e.g., different interleaving interval, different coding, and so on) based on the power control command. The R-SCH power control subchannel may also be partitioned into a number of "sub-subchannels", each of which may be used in any of the manners described above. The sub-subchannels may have the same or different bit rates. The remote terminal may apply the power control immediately upon receiving the command, or may apply the command at the next frame boundary.

[1090] The ability to reduce the R-SCH transmit power by a large amount (or down to zero) without terminating the communication session is especially advantageous to achieve better utilization of the reverse link. Temporary reduction or suspension of a packet data transmission can typically be tolerated by the remote terminal. These power control schemes can be advantageously used to reduce interference from a high rate remote terminal.

[1091] Power control of the R-SCH may be achieved in various manners. In one embodiment, a base station monitors the received power from the remote terminals with a power meter. The base station may even be able to determine the amount of power received from each channel (e.g., the R-FCH, R-DCCH, R-SCH, and so on). The base station is also able to determine the interference, some of which may be contributed by remote terminals not being served by this base station. Based on the collected information, the base station may adjust the transmit power of some or all remote terminals based on various factors.

For example, the power control may be based on the remote terminals' category of service, recent performance, recent throughput, and so on. The power control is performed in a manner to achieve the desired system goals.

[1092] Power control may be implemented in various manners. Example implementations are described in U.S Patent No. 5,485,486, entitled "METHOD AND APPARATUS FOR CONTROLLING TRANSMISSION POWER IN A CDMA CELLULAR MOBILE TELEPHONE SYSTEM," issued January 16, 1996, U.S Patent No. 5,822,318, entitled "METHOD AND APPARATUS FOR CONTROLLING POWER IN A VARIABLE RATE COMMUNICATION SYSTEM," issued October 13, 1998, and U.S Patent No. 6,137,840, entitled "METHOD AND APPARATUS FOR PERFORMING FAST POWER CONTROL IN A MOBILE COMMUNICATION SYSTEM," issued October 24, 2000, all assigned to the assignee of the present application and incorporated herein by reference.

[1093] In a typical method of power control that is used to control the level of the R-PICH channel, the base station measures the level of the R-PICH, compares it to a threshold, and then determines whether to increase or decrease the power of the remote terminal. The base station transmits a bit to the remote terminal instructing it to increase or decrease its output power. If the bit is received in error, the remote terminal will transmit at the incorrect power. During the next measurement of the R-PICH level received by the base station, the base station will determine that the received level is not at the desired level and send a bit to the remote terminal to change its transmit power. Thus, bit errors do not accumulate and the loop controlling the remote terminal's transmit power will stabilize to the correct value.

[1094] Errors in the bits sent to the remote terminal to control the traffic-to-pilot ratio for congestion power control can cause the traffic-to-pilot ratio to be other than that desired. However, the base station typically monitors the level of the R-PICH for reverse power control or for channel estimation. The base station can also monitor the level of the received R-SCH. By taking the ratio of the R-SCH level to the R-PICH level, the base station can estimate the traffic-to-pilot ratio in use by the remote terminal. If the traffic-to-pilot ratio is not that which is desired, then the base station can set the bit that controls the traffic-to-

pilot ratio to correct for the discrepancy. Thus, there is a self-correction for bit errors.

[1095] Once a remote terminal has received a grant for the R-SCH, the remote terminal typically transmits at the granted rate (or below in case it doesn't have enough data to send or does not have sufficient power) for the duration of the grant. The channel load from other remote terminals can vary quite quickly as a result of fading and the like. As such, it may be difficult for the base station to estimate the loading precisely in advance.

[1096] In an embodiment, a "congestion" power control subchannel may be provided to control a group of remote terminals in the same manner. In this case, instead of a single remote terminal monitoring the power control subchannel to control the R-SCH, a group of remote terminals monitor the control subchannel. This power control subchannel can be at 100 bps or at any other transmission rate. In one embodiment, the congestion control subchannel is implemented with the power control subchannel used for the R-SCH. In another embodiment, the congestion control subchannel is implemented as a "sub-subchannel" of the R-SCH power control subchannel. In yet another embodiment, the congestion control subchannel is implemented as a subchannel different from the R-SCH power control subchannel. Other implementations of the congestion control subchannel may also be contemplated and are within the scope of the invention.

[1097] The remote terminals in the group may have the same category service (e.g., remote terminals having low priority available bit rate services) and may be assigned to a single power control bit per base station. This group control based on a single power control stream performs similar to that directed to a single remote terminal to provide for congestion control on the reverse link. In case of capacity overload, the base station may direct this group of remote terminals to reduce their transmit power or their data rates, or to temporarily stop transmitting, based on a single control command. The reduction in the R-SCH transmit power in response to the congestion control command may be a large downward step relative to the transmit power of the pilot channel.

[1098] The advantage of a power control stream going to a group of remote terminals instead of a single remote terminal is that less overhead power is

required on the forward link to support the power control stream. It should be noted that the transmit power of a bit in the power control stream can be equal to the power of the normal power control stream used to the control the pilot channel for the remote terminal that requires the most power. That is, the base station can determine the remote terminal in the group that requires the greatest power in its normal power control stream and then use this power to transmit the power control bit used for congestion control.

[1099] FIG. 7 is a flow diagram that illustrates a variable rate data transmission on the R-SCH with fast congestion control, in accordance with an embodiment of the invention. During the transmission on the R-SCH, the remote terminal transmits in accordance with the data rate granted in the *Reverse Supplemental Channel Assignment Mini Message* (RSAMM). If variable rate operation is permitted on the R-SCH, the remote terminal may transmit at any one of a number of permitted data rates.

[1100] If the remote terminal's R-SCH has been assigned to a congestion control subchannel, then, in an embodiment, the remote terminal adjusts the traffic-to-pilot ratio based upon the bits received in the congestion control subchannel. If variable rate operation is permitted on the R-SCH, the remote terminal checks the current traffic-to-pilot ratio. If it is below the level for a lower data rate, then the remote terminal reduces its transmission rate to the lower rate. If it is equal to or above the level for a higher data rate, then the remote terminal increases its transmission rate to the higher rate if it has sufficient data to send.

[1101] Prior to the start of each frame, the remote terminal determines the rate to use for transmitting the next data frame. Initially, the remote terminal determines whether the R-SCH traffic-to-pilot ratio is below that for the next lower rate plus a margin Δ_{low}, at step 712. If the answer is yes, a determination is made whether the service configuration allows for a reduction in the data rate, at step 714. And if the answer is also yes, the data rate is decreased, and the same traffic-to-pilot ratio is used, at step 716. And if the service configuration does not allow for a rate reduction, a particular embodiment would permit the remote terminal to temporarily stop transmitting.

[1102] Back at step 712, if the R-SCH traffic-to-pilot ratio is not above that for the next lower data rate plus the margin Δ_{low}, a determination is next made as to whether the R-SCH traffic-to-pilot ratio is greater than that for the next higher data rate minus a margin Δ_{high}, at step 718. If the answer is yes, a determination is made whether the service configuration allows for an increase in the data rate, at step 720. And if the answer is also yes, the transmission rate is increased, and the same traffic-to-pilot ratio is used, at step 722. And if the service configuration does not allow for a rate increase, the remote terminal transmits at the current rate.

[1103] FIG. 8 is a diagram illustrating improvement that may be possible with fast control of the R-SCH. On the left frame, without any fast control of the R-SCH, the rise-over-thermal at the base station varies more widely, exceeding the desired rise-over-thermal level by a larger amount in some instances (which may result in performance degradation for the data transmissions from the remote terminals), and falling under desired rise-over-thermal level by a larger amount in some other instances (resulting in under-utilization of the reverse link resources). In contrast, on the right frame, with fast control of the R-SCH, the rise-over-thermal at the base station is maintained more closely to the desired rise-over-thermal level, which results in improved reverse link utilization and performance.

[1104] In an embodiment, a base station may schedule more than one remote terminal (via SCAM or ESCAM) to transmit, in response to receiving multiple requests (via SCRM or SCRMM) from different remote terminals. The granted remote terminals may thereafter transmit on the R-SCH. If overloading is detected at the base station, a "fast reduce" bit stream may be used to turn off (i.e., disable) a set of remote terminals (e.g., all except one remote terminal). Alternatively, the fast reduce bit stream may be used to reduce the data rates of the remote terminals (e.g., by half). Temporarily disabling or reducing the data rates on the R-SCH for a number of remote terminals may be used for congestion control, as described in further detail below. The fast reduce capability may also be advantageously used to shorten the scheduling delay.

[1105] When the remote terminals are not in soft handoff with other base stations, the decision on which remote terminal is the most advantaged (efficient) to use the reverse link capacity may be made at the BTS. The most efficient remote terminal may then be allowed to transmit while the others are temporarily disabled. If the remote terminal signals the end of its available data, or possibly when some other remote terminal becomes more efficient, the active remote terminal can quickly be changed. These schemes may increase the throughput of the reverse link.

[1106] In contrast, for a usual set up in a cdma2000 system, a R-SCH transmission can only start or stop via layer 3 messaging, which may take several frames from composing to decoding at the remote terminal to get across. This longer delay causes a scheduler (e.g., at the base station or BSC) to work with (1) less reliable, longer-term predictions about the efficiency of the remote terminal's channel condition (e.g., the reverse link target pilot Ec/(No+Io) or set point), or (2) gaps in the reverse link utilization when a remote terminal notifies the base station of the end of its data (a common occurrence since a remote terminal often claims it has a large amount of data to send to the base station when requesting the R-SCH).

[1107] Referring back to FIG. 2, the elements of remote terminal 106 and base station 104 may be designed to implement various aspects of the invention, as described above. The elements of the remote terminal or base station may be implemented with a digital signal processor (DSP), an application specific integrated circuit (ASIC), a processor, a microprocessor, a controller, a microcontroller, a field programmable gate array (FPGA), a programmable logic device, other electronic units, or any combination thereof. Some of the functions and processing described herein may also be implemented with software executed on a processor, such as controller 230 or 270.

[1108] Headings are used herein to serve as general indications of the materials being disclosed, and are not intended to be construed as to scope.

[1109] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those

skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

[1110] WHAT IS CLAIMED IS:

CLAIMS

1. A channel structure capable of supporting data transmission on a reverse link of a wireless communication system, comprising:

a reverse fundamental channel configurable to transmit data and signaling on the reverse link;

a reverse supplemental channel assignable and configurable to transmitted packet data on the reverse link;

a reverse control channel configurable to transmit signaling on the reverse link; and

a forward power control channel configurable to transmit first and second power control streams for the reverse link for a particular remote terminal, wherein

the first power control stream is used to control the transmit power of the reverse supplemental channel in combination with at least one other reverse link channel, and

the second power control stream is used to control a transmit characteristic of the reverse supplemental channel.

2. The channel structure of claim 1, wherein the second power control stream is used to control the transmit power of the reverse supplemental channel relative to that of a designated reverse link channel.

3. The channel structure of claim 1, wherein the second power control stream is used to control the data rate of the reverse supplemental channel.

4. The channel structure of claim 1, further comprising:

a forward acknowledgment channel configurable to transmit, on the forward link, signaling indicative of received status of the packet data transmission on the reverse link.

5. The channel structure of claim 4, wherein the forward acknowledgment channel is configurable to transmit an acknowledgment or a negative acknowledgment for each transmitted data frame on the reverse supplemental channel.

6. The channel structure of claim 5, wherein the acknowledgment or negative acknowledgment for each transmitted data frame is transmitted a plurality of times on the forward acknowledgment channel.

7. The channel structure of claim 1, wherein the reverse control channel is configurable to transmit signaling used to assign and de-assign the reverse supplemental channel.

8. The channel structure of claim 1, further comprising:

a reverse rate indicator channel configurable to transmit on the reverse link information related to a packet data transmission on the reverse link.

9. A channel structure capable of supporting data transmission on a reverse link of a wireless communication system, comprising:

a reverse fundamental channel configurable to transmit data and signaling on the reverse link;

a reverse supplemental channel assignable and configurable to transmitted packet data on the reverse link;

a reverse control channel configurable to transmit signaling on the reverse link; and

a forward power control channel configurable to transmit first and second power control streams for the reverse link for a particular remote terminal, wherein

the first power control stream is used to control the transmit power of the reverse supplemental channel in combination with at least one other reverse link channel, and

the second power control stream is configured to control a transmit characteristic of a group of remote terminals.

10. The channel structure of claim 9, wherein the second power control stream is used to similarly control the transmit power or data rate of the group of remote terminals.

11. The channel structure of claim 9, wherein the second power control stream is used to enable and disable transmissions on reverse supplemental channels assigned to the group of remote terminals.

12. A method for transmitting data on a reverse link of a wireless communication system, comprising:

transmitting a frame of data on the reverse link via a data channel;

temporarily retaining the data frame in a buffer;

monitoring for a message on a forward link indicating a received status of the transmitted data frame; and

processing the data frame based on the received message.

13. The method of claim 12, wherein the processing includes;

retransmitting the data frame if the message indicates that the transmitted data frame was incorrectly received.

14. The method of claim 12, wherein the processing includes;

discarding the data frame from the buffer if the message indicates that the transmitted data frame was correctly received.

15. The method of claim 12, wherein the processing includes;

retaining the data frame in the buffer if the message is not properly detected.

16. The method of claim 12, further comprising:

monitoring for a second transmission of the message;

wherein the processing of the data frame is based on one or more received messages for the data frame.

17. The method of claim 16, further comprising:
combining the received messages for the data frame to provide a more reliable message.

18. The method of claim 12, further comprising:
identifying the transmitted data frame with a sequence number.

19. The method of claim 18, further comprising:
transmitting the sequence number of the transmitted data frame via a signaling channel.

20. The method of claim 12, further comprising:
identifying the transmitted data frame as either a first transmission or a retransmission.

21. A method for transmitting data on a reverse link of a wireless communication system, comprising:
transmitting a frame of data on the reverse link via a data channel;
temporarily retaining the data frame in a buffer;
monitoring for a message on a forward link indicating a received status of the transmitted data frame;
retransmitting the data frame if the message indicates that the transmitted data frame was incorrectly received;
discarding the data frame from the buffer if the message indicates that the transmitted data frame was correctly received; and
retaining the data frame in the buffer if the message is not properly detected.

22. A method for controlling transmit power of a supplemental channel in a reverse link of a wireless communication system, comprising:

receiving a first power control stream for controlling the transmit power of the supplemental channel in combination with at least one other reverse link channel;

receiving a second power control stream for controlling a transmit characteristic of the supplemental channel; and

adjusting the transmit power and characteristic of the supplemental channel based on the first and second power control streams.

23. The method of claim 22, wherein the second power control stream controls the transmit power of the supplemental channel relative to that of a designated reverse link channel.

24. The method of claim 22, wherein the second power control stream controls a data rate of the supplemental channel.

25. The method of claim 22, wherein the second power control stream enables and disables transmission on the supplemental channel.

26. The method of claim 22, wherein the transmit power of the supplemental channel is adjusted by a larger step in response to the second power control stream than for the first power control stream.

27. The method of claim 22, wherein the second power control stream is assigned to a plurality of remote terminals.

28. The method of claim 28, wherein supplemental channels for the plurality of remote terminals are controlled in similar manner by the second power control stream.

29. A remote terminal in a wireless communication system, comprising:

a transmit data processor configurable to process and transmit

data and signaling on a reverse fundamental channel,

packet data on an assigned reverse supplemental channel,

signaling on a reverse control channel, and

information related to a packet data transmission on a reverse indicator channel;

a receive data processor configurable to receive a plurality of power control streams on a forward power control channel; and

a controller operatively coupled to the transmit and receive data processors and configured to control one or more transmit characteristics of the reverse supplemental channel based on the plurality of power control streams.

30. The remote terminal of claim 29, wherein the receive data processor is further configurable to receive, on a forward acknowledgment channel, signaling indicative of received status of a packet data transmission on the reverse supplemental channel.

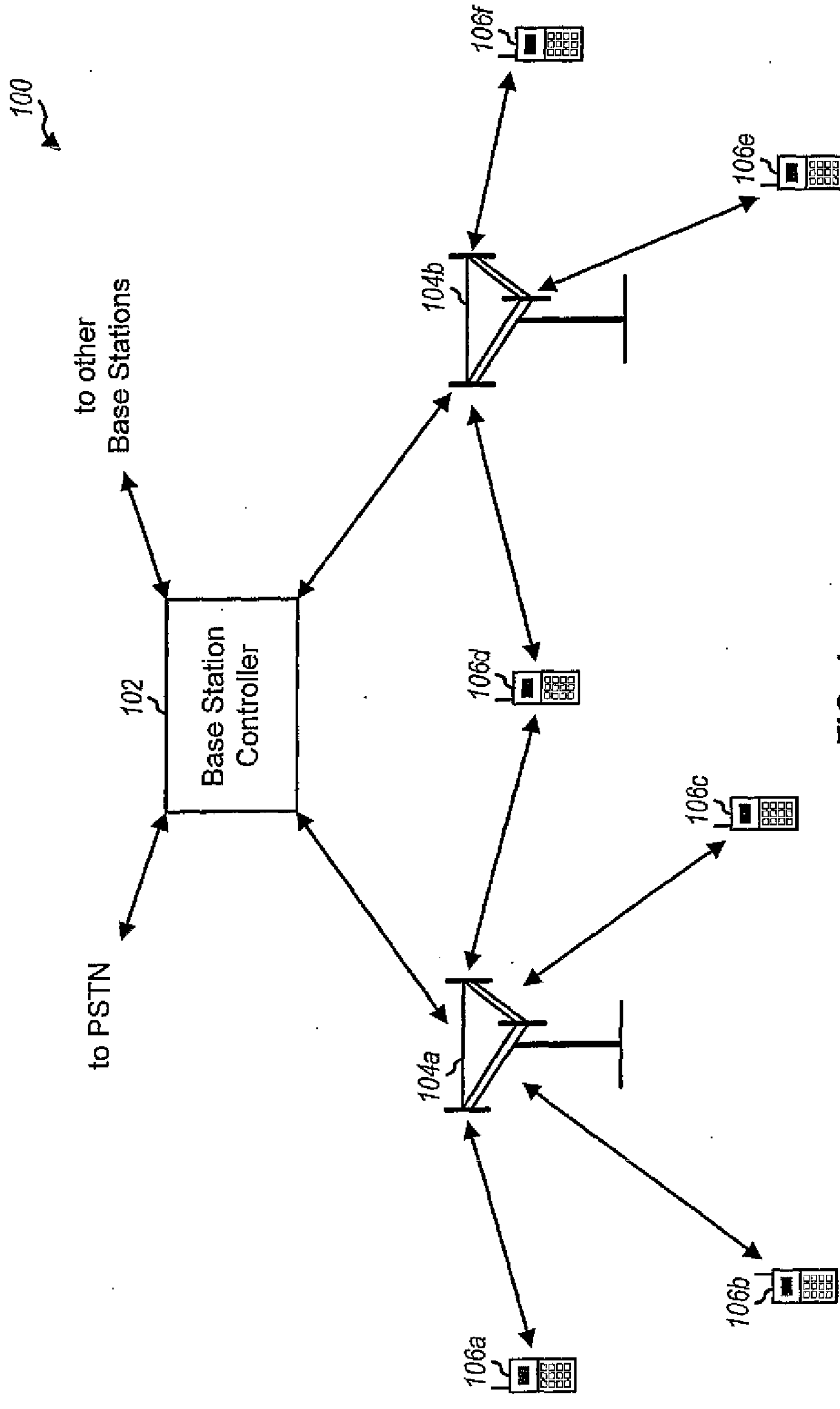

100
to other Base Stations
to PSTN
102
Base Station Controller
104a
104b
106a
106b
106c
106d
106e
106f

FIG. 1

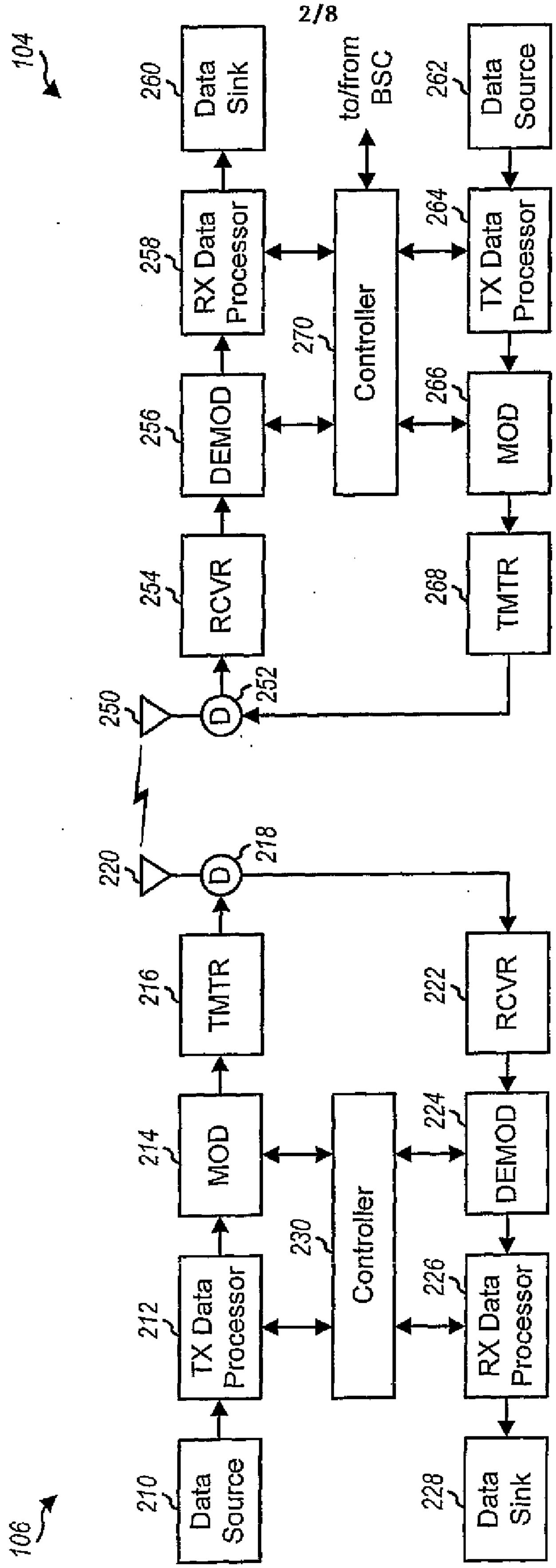

104
260
Data Sink
258
RX Data Processor
256
DEMOD
254
RCVR
250
252
D
to/from BSC
270
Controller
262
Data Source
264
TX Data Processor
266
MOD
268
TMTR
220
218
D
216
TMTR
214
MOD
212
TX Data Processor
210
Data Source
106
230
Controller
222
RCVR
224
DEMOD
226
RX Data Processor
228
Data Sink

FIG. 2

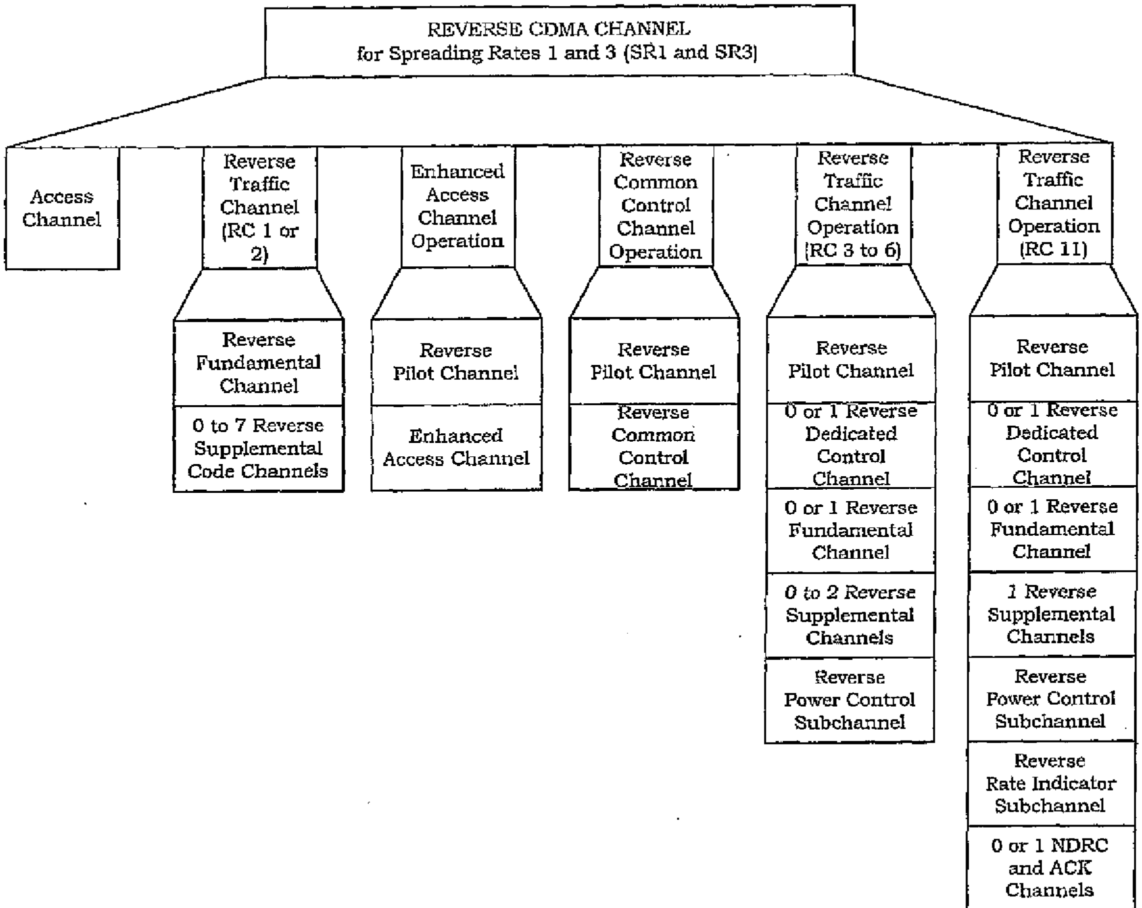

REVERSE CDMA CHANNEL for Spreading Rates 1 and 3 (SR1 and SR3)
Access Channel
Reverse Traffic Channel (RC 1 or 2)
Enhanced Access Channel Operation
Reverse Common Control Channel Operation
Reverse Traffic Channel Operation (RC 3 to 6)
Reverse Traffic Channel Operation (RC 11)
Reverse Fundamental Channel
0 to 7 Reverse Supplemental Code Channels
Reverse Pilot Channel
Enhanced Access Channel
Reverse Pilot Channel
Reverse Common Control Channel
Reverse Pilot Channel
0 or 1 Reverse Dedicated Control Channel
0 or 1 Reverse Fundamental Channel
0 to 2 Reverse Supplemental Channels
Reverse Power Control Subchannel
Reverse Pilot Channel
0 or 1 Reverse Dedicated Control Channel
0 or 1 Reverse Fundamental Channel
1 Reverse Supplemental Channels
Reverse Power Control Subchannel
Reverse Rate Indicator Subchannel
0 or 1 NDRC and ACK Channels

FIG. 3A

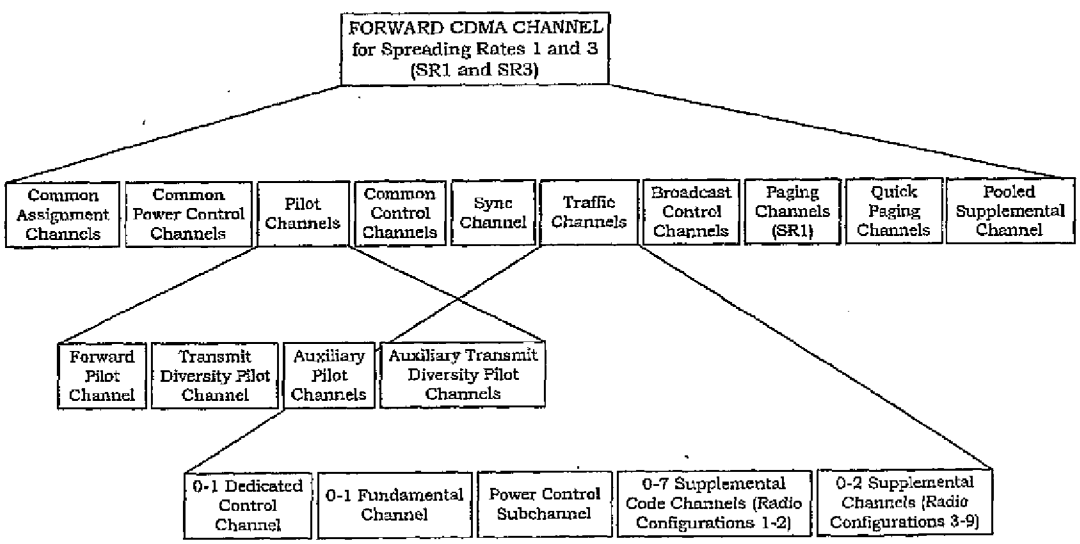

FORWARD CDMA CHANNEL for Spreading Rates 1 and 3 (SR1 and SR3)
Common Assignment Channels
Common Power Control Channels
Pilot Channels
Common Control Channels
Sync Channel
Traffic Channels
Broadcast Control Channels
Paging Channels (SR1)
Quick Paging Channels
Pooled Supplemental Channel
Forward Pilot Channel
Transmit Diversity Pilot Channel
Auxiliary Pilot Channels
Auxiliary Transmit Diversity Pilot Channels
0-1 Dedicated Control Channel
0-1 Fundamental Channel
Power Control Subchannel
0-7 Supplemental Code Channels (Radio Configurations 1-2)
0-2 Supplemental Channels (Radio Configurations 3-9)

FIG. 3B

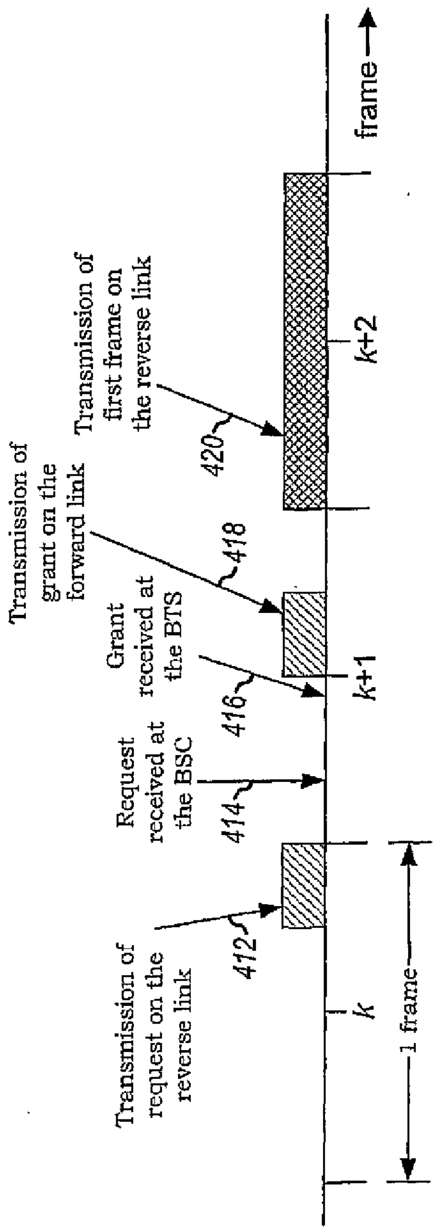
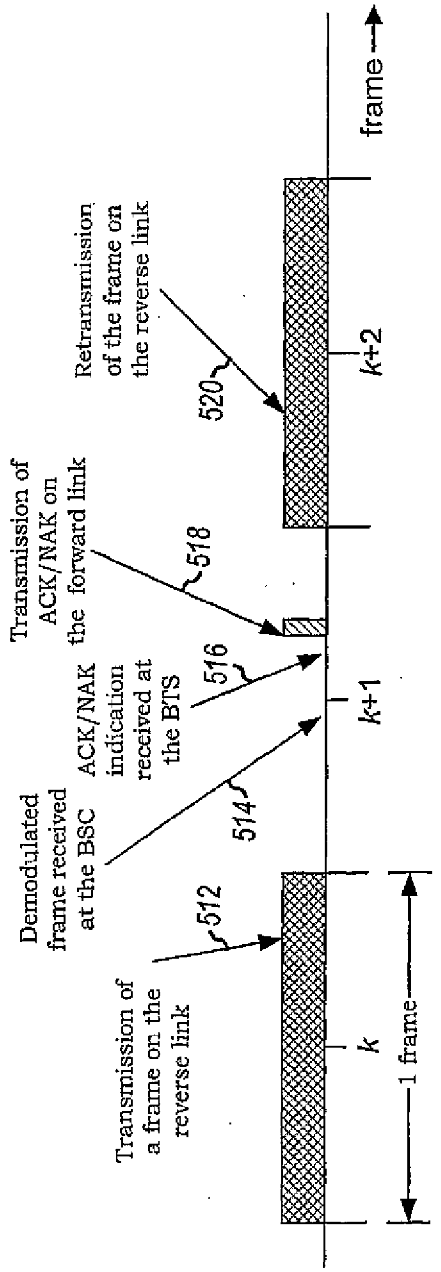
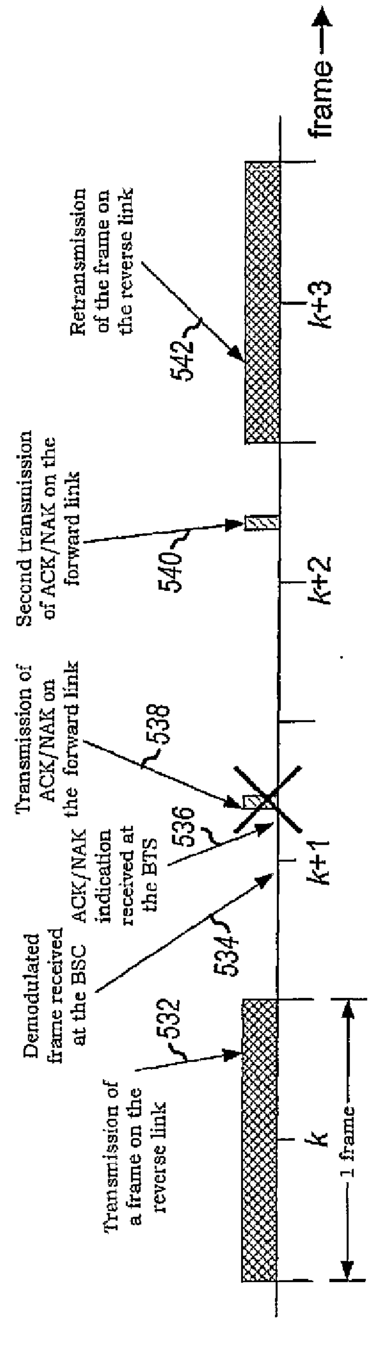

Transmission of request on the reverse link
412
Request received at the BSC
414
Grant received at the BTS
416
Transmission of grant on the forward link
418
Transmission of first frame on the reverse link
420
k
k+1
k+2
frame
1 frame

FIG. 4



Transmission of a frame on the reverse link
512
Demodulated frame received at the BSC
514
ACK/NAK indication received at the BTS
516
Transmission of ACK/NAK on the forward link
518
Retransmission of the frame on the reverse link
520
k
k+1
k+2
1 frame
frame

FIG. 5A



Transmission of a frame on the reverse link
532
Demodulated frame received at the BSC
534
ACK/NAK indication received at the BTS
536
Transmission of ACK/NAK on the forward link
538
Second transmission of ACK/NAK on the forward link
540
Retransmission of the frame on the reverse link
542
k
k+1
k+2
k+3
1 frame
frame

FIG. 5B

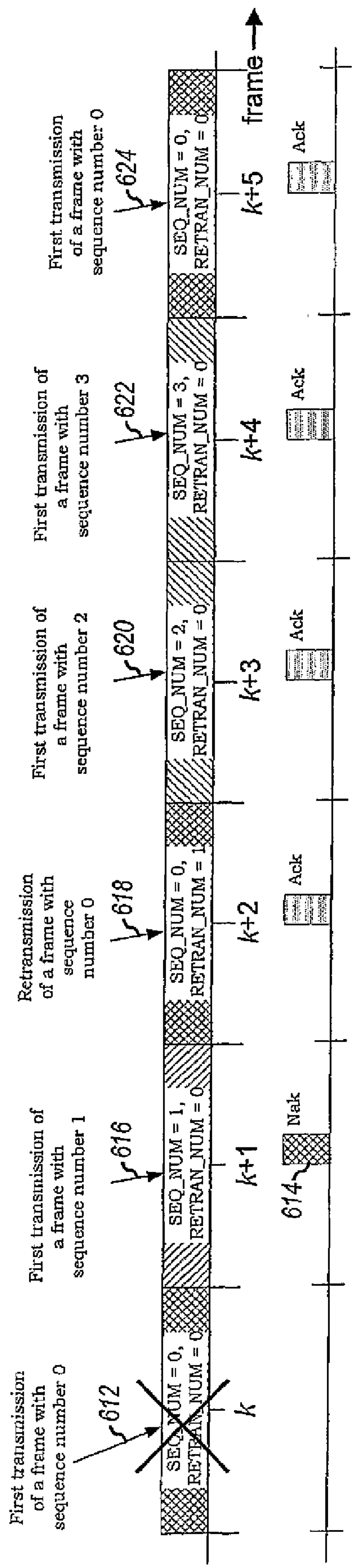

First transmission of a frame with sequence number 0
612
SEQ_NUM = 0, RETRAN_NUM = 0
k
First transmission of a frame with sequence number 1
616
SEQ_NUM = 1, RETRAN_NUM = 0
k+1
614
Nak
Retransmission of a frame with sequence number 0
618
SEQ_NUM = 0, RETRAN_NUM = 1
k+2
Ack
First transmission of a frame with sequence number 2
620
SEQ_NUM = 2, RETRAN_NUM = 0
k+3
Ack
First transmission of a frame with sequence number 3
622
SEQ_NUM = 3, RETRAN_NUM = 0
k+4
Ack
First transmission of a frame with sequence number 0
624
SEQ_NUM = 0, RETRAN_NUM = 0
k+5
Ack
frame

FIG. 6A

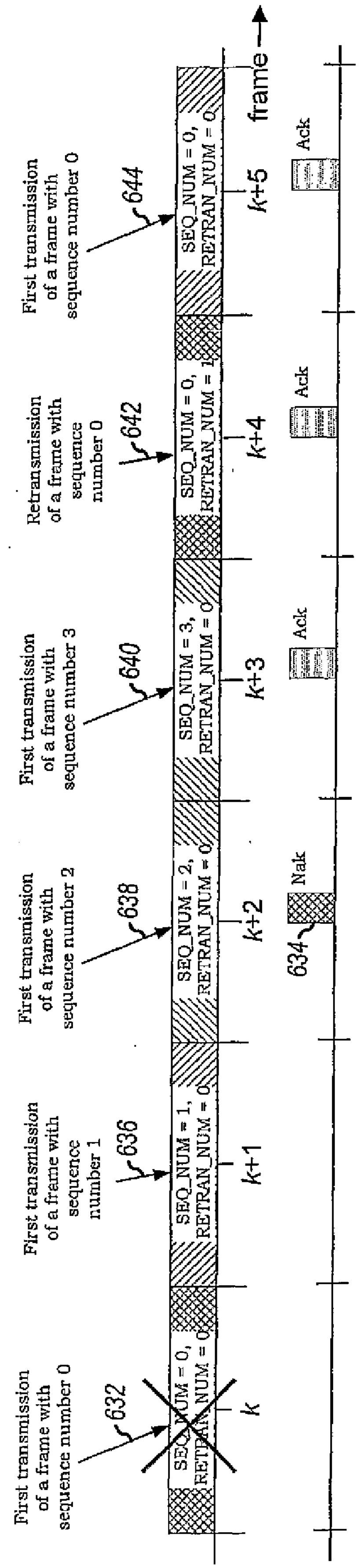

First transmission of a frame with sequence number 0
632
SEQ_NUM = 0, RETRAN_NUM = 0
k
First transmission of a frame with sequence number 1
636
SEQ_NUM = 1, RETRAN_NUM = 0
k+1
First transmission of a frame with sequence number 2
638
SEQ_NUM = 2, RETRAN_NUM = 0
k+2
634
Nak
First transmission of a frame with sequence number 3
640
SEQ_NUM = 3, RETRAN_NUM = 0
k+3
Ack
Retransmission of a frame with sequence number 0
642
SEQ_NUM = 0, RETRAN_NUM = 1
k+4
Ack
First transmission of a frame with sequence number 0
644
SEQ_NUM = 0, RETRAN_NUM = 0
k+5
Ack
frame

FIG. 6B

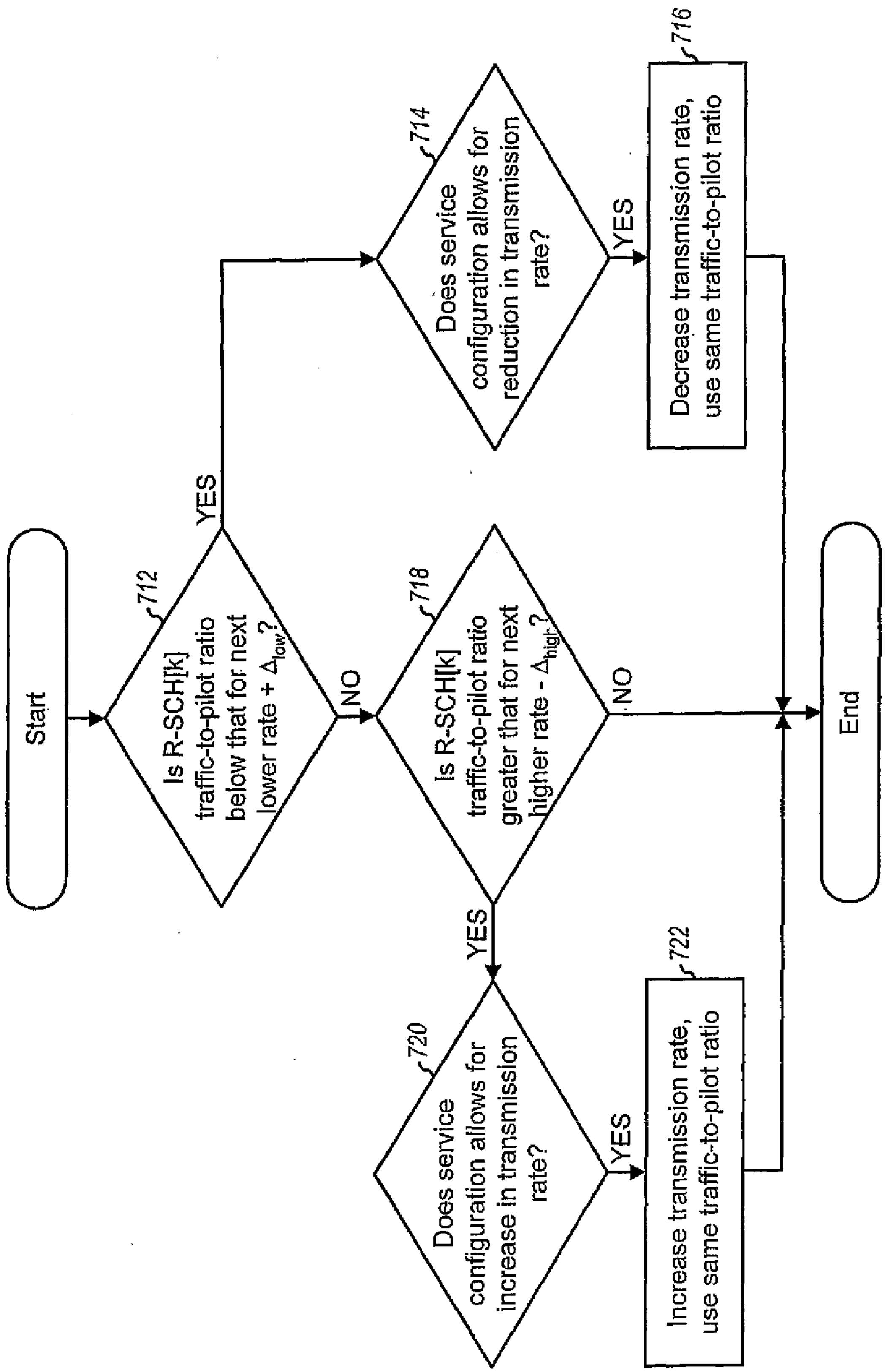

Start
712
Is R-SCH[k] traffic-to-pilot ratio below that for next lower rate + Δ_{low}?
YES
NO
714
Does service configuration allows for reduction in transmission rate?
YES
716
Decrease transmission rate, use same traffic-to-pilot ratio
718
Is R-SCH[k] traffic-to-pilot ratio greater that for next higher rate - Δ_{high}?
NO
YES
720
Does service configuration allows for increase in transmission rate?
YES
722
Increase transmission rate, use same traffic-to-pilot ratio
End

FIG. 7

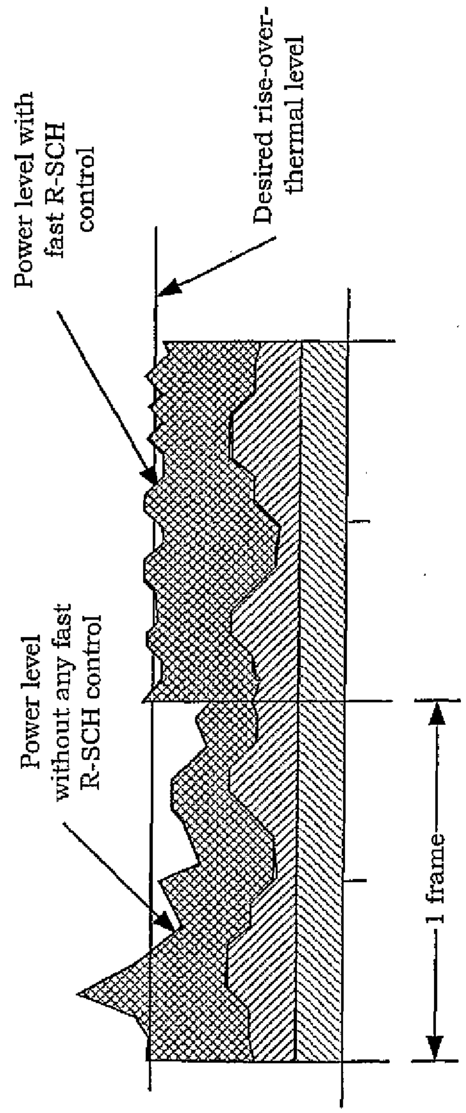

Power level with fast R-SCH control
Desired rise-over-thermal level
Power level without any fast R-SCH control
1 frame

FIG. 8

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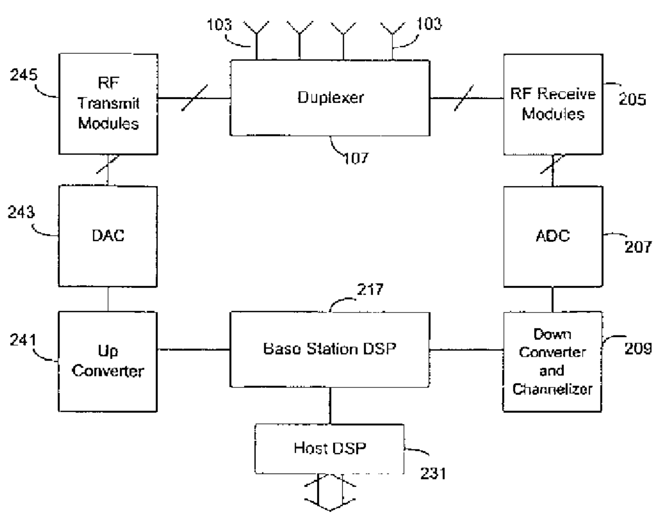
(54) Title: FREQUENCY DEPENDENT CALIBRATION OF A WIDEBAND RADIO SYSTEM USING NARROWBAND CHANNELS



(57) Abstract: A method and apparatus are provided that determine group delay for a set of transmit or receive chains over a wide frequency band without causing significant interference with simultaneous users of the system. In one embodiment, the invention includes an antenna array adapted to transmit and receive radio communications signals with a plurality of other terminals, the communications signals each using a particular minimum bandwidth, a transmit chain to transmit a calibration signal through the antenna array to a transponder on at least two different frequency bands within the minimum bandwidth, and a receive chain to receive through the antenna array a transponder signal from the transponder, the transponder signal being received on at least two different frequency bands and being based on the calibration signal. A signal processor determines a frequency dependent calibration vector based on the at least two frequency bands of the transponder signal as received through the receive chain.

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For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

FREQUENCY DEPENDENT CALIBRATION OF A WIDEBAND RADIO SYSTEM USING NARROWBAND CHANNELS

BACKGROUND OF THE INVENTION

Field of the Invention

The invention relates generally to the field of digital signal communications and to receive and transmit chain calibration. More particularly, the invention relates to calibrating the group delay using narrowband signals at more than one frequency.

Description of the Related Art

Radio communications capacity can be greatly increased using directional, rather than omni-directional radio transmission. One way to transmit directional signals and directionally receive signals is by using beam forming and nulling through an array of antennas. The precision of the beam forming and nulling through the antenna array, can be improved if the transmit and receive chains are both calibrated. Calibration can be applied to the chain from the digital interface at baseband to the field radiated from or received at each antenna element. One way of making the calibration is to have a transponder separated from the antenna array listen to the output of the antenna array on a base station downlink frequency. The transponder receives a downlink calibration signal from the base station and then re-transmits it on an uplink frequency. By selecting appropriate signals to transmit and appropriate signals to receive, the base station can apply signal processing to estimate compensations in phase and amplitude to calibrate its transmit and receive chains.

A remote transponder calibration system is shown, for example, in U.S. Patent No. 5,546,090 to Roy, III et al. That patent describes calibrating a narrowband FDD (frequency division duplex) system for phase and amplitude at each transmit and receive chain. In an FDD system, unused time and frequency slots typically occur on occasion and these can be used to send and receive a narrowband calibration signal. In a typical spread spectrum system, however, there are no unused time and frequency slots to use for calibration. A spread spectrum system, for example a CDMA (code division multiple access) system, as opposed to FDMA (frequency division multiple access) and TDMA (time division multiple access) systems, has multiple users using the same radio channel at the same time. If the transponder is designed to receive and transmit the signal using the same spread spectrum channel that is used for traffic, then the additional energy added to the channel by calibration will reduce system

capacity. A typical transponder will receive all of the downlink traffic including the calibration signal, shift the frequency, amplify it and send all of the traffic back to the base station. This results in a very large amount of energy being sent by the transponder on the uplink and may effectively overpower all other traffic. As a result, calibration will affect both the downlink and uplink channel capacity. For calibrating group delay for a set of transmitters or receivers, a calibration signal normally is transmitted across a wider band of frequencies further ensuring interruptions to normal traffic.

BRIEF SUMMARY OF THE INVENTION

A method and apparatus are provided that determine group delay for a set of transmit or receive chains over a wide frequency band without causing significant interference with simultaneous users of the system. In one embodiment, the invention includes an antenna array adapted to transmit and receive radio communications signals with a plurality of other terminals, the communications signals each using a particular minimum bandwidth, a transmit chain to transmit a calibration signal through the antenna array to a transponder on at least two different frequency bands within the minimum bandwidth, and a receive chain to receive through the antenna array a transponder signal from the transponder, the transponder signal being received on at least two different frequency bands and being based on the calibration signal. A signal processor determines a frequency dependent calibration vector based on the at least two frequency bands of the transponder signal as received through the receive chain.

Other features of the present invention will be apparent from the accompanying drawings and from the detailed description that follows.

BRIEF DESCRIPTION OF THE SEVERAL VIEWS OF THE DRAWINGS

The present invention is illustrated by way of example, and not by way of limitation, in the figures of the accompanying drawings in which like reference numerals refer to similar elements and in which:

Figure 1 is a block diagram illustrating an exemplary architecture of a wireless communication system base station appropriate for use with one embodiment of the present invention;

Figure 2 is a block diagram illustrating an exemplary architecture of a wireless transponder system appropriate for use with the base station of Figure 1;

Figure 3 is a process flow diagram showing the calibration of a receive chain; and

Figure 4 is a process flow diagram showing the calibration of a transmit chain.

DETAILED DESCRIPTION OF THE INVENTION

Introduction

In one embodiment, the present invention includes a method for calibrating the group delay of multiple transmit and receive chains of a wideband adaptive antenna base station using a narrowband transponder. In order to calibrate the group delay of the transmit and the receive chains, the base station transmits a different narrowband calibration signal over each of the transmit chains on at least two different frequencies in the downlink frequency band. These signals are then received by the narrowband transponder and retransmitted to the base station as narrowband signals in the wideband uplink frequency band. In this application, the radios in the adaptive antenna base station support wideband channels. However, in order to avoid creating any unnecessary interference, the calibration signals and the transponder signals are narrowband. In other words, the calibration signals occupy only narrow portions of the wideband channel. The transponder only receives in these narrow frequency bands and only retransmits the signals in correspondingly narrow portions of the uplink band.

Since the narrowband signals add only a small amount of energy to the wideband uplink and downlink channels, the calibration can be done while regular data traffic is being supported by the base station. The narrower the bandwidth of the calibration signals, the less will be the amount of energy that will be added to the system. For wideband spread spectrum systems the narrowband signals can easily be one tenth, or one hundredth as wide as the regular data traffic channels. For frequency division systems, the narrowband signals can still be one third to one fifth the width of the traffic channels. Proper selection of the signal power levels can further reduce the impact on regular traffic. Using multiple narrowband signals and transponder bands it is possible to calibrate for more general phase and gain variations as a function of

frequency. In a CDMA (Code Division Multiple Access) system, it is possible to design the CDMA system to be particularly insensitive to narrowband signals.

In one embodiment, the transponder only receives and re-transmits on narrow bands within the traffic bands of the wider band system to be calibrated. The system can have a set of wideband transmitters with antenna elements and a set of wideband receivers with antenna elements or a single set of elements can be common to the transmitters and the receivers. In both cases, system performance is normally improved with frequent calibration of the group delay for both the transmit chain and the receive chain. The group delay calibration vectors can be different for the receive chain and the transmit chain. In one example, the system has a multi-channel base station that communicates with multiple subscribers up to 10km away using CDMA with SDMA (spatial division multiple access). For this system, it has been found that calibrations every hour or two will noticeably improve performance. With such frequent calibrations, the impact of calibration on normal operations can be important. According to the present invention, the impact of calibration on normal operations can be minimized with a narrowband calibration transponder.

On each narrow frequency calibration band, different signals can be transmitted through two or more transmit chains. The signals can be differentiated, for example, by modulating different sequences onto the signals. In one embodiment, the sequences are orthogonal sequences to aid in demodulation. In another embodiment, the sequences are modulated onto the signals as spreading codes. This allows de-spreading codes to be used on the received signal so that the signal from each transmit chain can be distinguished. The transponder receives these signals and re-transmits them in the base station uplink band. The signals received by the base station can then be processed in order to measure any desired relative characteristics of the signals. For example, the signals can be used to find the relative phase and amplitude of the involved transmit chains and the relative phase and amplitude of all the receive chains. By transmitting different signals over the different transmit chains, the signals can be differentiated when received. This allows characteristics such as relative phase and amplitude to be estimated separately for each transmit chain. The characteristics can be used to determine spatial signatures for the uplink and downlink as well as to calculate frequency dependent calibration vectors. Combining phase measurements at different frequencies, a group delay calibration vector can be derived.

The relative phase and amplitude of the transmit chains can be estimated by receiving the different signals at a single antenna and then estimating the channel for each of the different signals transmitted over the different transmit chains. The relative phase and amplitude of the receive chains can be estimated by transmitting a single calibration signal over a single transmit chain and receiving it over the different receive chains. The channel received over each receive chain can then be estimated and compared to find spatial signatures and for calibration. As a result, if the calibration signal is sent once over all transmit chains and then the corresponding transponder signal is received through all receive chains, the entire array can be calibrated based on a single downlink and uplink burst. Since the transmit and receive calibration vector determinations need not be coupled to each other, performing both calibrations on the same burst increases efficiency and reduces the effects on traffic. If the calibration signal is transmitted on two or more different frequencies either at the same time or at different times close together, then the group delay can be derived.

As an alternative, just a few or even two of the transmit or receive chains can be calibrated at one time. If all the transmit or receive chains are not involved in each calibration measurement, then repeated calibration measurements with different sets of transmit or receive chains can be performed so that all relative phases and amplitudes can be measured among all the transmit and receive antennas. Accuracy is improved if there is a common transmit or receive chain in each of the measurements. This allows the measured phases and amplitudes to be related to each other with reference to the common chain. Typically, one of the receive chains is designated as a reference receive chain and calibration signals are measured in pairs with each receive chain being paired with the reference chain. Since the reference chain participates in every measurement, all of the other chains can be referenced to each other through the reference chain. After the receive chains are calibrated, a similar process is performed with the transmit chains being measured in pairs against the reference. It is not important which particular chain is selected to be the reference and the receive and transmit references need not have any relationship to each other. The calibration vectors can be expressed as variations from the reference or from any arbitrary standard such as an average, mean, or median of the differences between the receive or transmit chains, respectively.

In one embodiment, the present invention is implemented in an SDMA radio data communications system. In such a spatial division system, each terminal is

associated with a set of spatial parameters that relate to the radio communications channel between, for example, the base station and a user terminal. The spatial parameters comprise a spatial signature for each terminal. Using the spatial signature and arrayed antennas, the RF energy from the base station can be more precisely directed at a single user terminal, reducing interference with and lowering the noise threshold for other user terminals. Conversely, data received from several different user terminals at the same time can be resolved at lower receive energy levels. With spatial division antennas at the user terminals, the RF energy required for communications can be even less. The benefits are even greater for subscribers that are spatially separated from one another. The spatial signatures can include such things as the spatial location of the transmitters, the directions-of-arrival (DOAs), times-of-arrival (TOAs) and the distance from the base station.

Estimates of parameters such as signal power levels, DOAs, and TOAs can be determined using known training sequences placed in digital data streams for the purpose of channel equalization in conjunction with sensor (antenna) array information. This information is then used to calculate appropriate weights for spatial demultiplexers, multiplexers, and combiners. Extended Kalman filters or other types of linear filters, well known in the art, can be used to exploit the properties of the training sequences in determining spatial parameters. Further details regarding the use of spatial division and SDMA systems are described, for example, in U.S. Patents Nos. 5,828,658, issued Oct. 27, 1998 to Ottersten et al. and 5,642,353, issued June 24, 1997 to Roy, III et al.

Base Station Structure

The present invention relates to wireless communication systems and may be a fixed-access or mobile-access wireless network. It may use spatial division technology in combination with wideband multiple access systems, such as code division multiple access (CDMA), and other spread spectrum type systems. Figure 1 shows an example of a base station of a wireless communications system or network suitable for implementing the present invention. The system or network includes a number of subscriber stations, also referred to as remote terminals or user terminals, (not shown). The base station may be connected to a wide area network (WAN) through its host DSP 231 for providing any required data services and connections external to the immediate wireless system. To support spatial division, a plurality of

antennas 103 is used, for example four antennas, although other numbers of antennas may be selected.

The outputs of the antennas are connected to a duplexer switch 107, which in this CDMA system is a frequency switch. Alternatively, separate transmit and receive antenna arrays can be used, in which case the duplexer is not necessary. When receiving, the antenna outputs are connected via the switch 107 to RF (radio frequency) receive modules 205, and are mixed down and channelized in a down converter 207. The down converted signals are then sampled and converted to digital in an ADC (analog to digital converter) 209. This can be done using FIR (finite impulse response) filtering techniques. The invention can be adapted to suit a wide variety of RF and IF (intermediate frequency) carrier frequencies and bands.

There are, in the present example, four antenna channel outputs, one from each antenna receive module 205. The particular number of channels can be varied to suit network needs. For each of the four receive antenna channels, the four down-converted outputs from the four antennas are fed to a digital signal processor (DSP) device 217 for further processing, including calibration. According to one aspect of this invention, four Motorola DSP56300 Family DSPs can be used as channel processors, one per receive channel. The timeslot processors 217 monitor the received signal power and estimate the phase and time alignment. They also determine smart antenna weights for each antenna element. These are used in the spatial division multiple access scheme to determine a signal from a particular remote user and to demodulate the determined signal.

The output of the channel processors 217 is demodulated burst data. This data is sent to the host DSP 231 whose main function is to control all elements of the system and interface with the higher level processing. The higher level processing provides the signals required for communications in all the different control and service communication channels defined in the system's communication protocols. The host DSP 231 can be a Motorola DSP56300 Family DSP. In addition, channel processors send the determined receive weights for each user terminal to the host DSP 231.

The host DSP 231 maintains state and timing information, receives uplink burst data from the channel processors 217, and programs the channel processors 217. In addition, it decrypts, descrambles, checks error detecting code, and deconstructs bursts of the uplink signals, then formats the uplink signals to be sent for higher level processing in other parts of the base station. With respect to the other parts of the

base station, it formats service data and traffic data for further higher processing in the base station, receives downlink messages and traffic data from the other parts of the base station, processes the downlink bursts and formats and sends the downlink bursts to the transmit chain, discussed below.

Transmit data from the host DSP 231 is used to produce analog transmit outputs which are sent to the RF transmitter (tx) modules 245. Specifically, the received data bits are converted via a DAC (digital to analog converter) 241 to analog transmit waveforms and up-converted into a complex modulated signal, at an IF frequency in an upconverter 243. The analog waveforms are sent to the transmit modules 245. The transmit modules 245 up-convert the signals to the transmission frequency and amplify the signals. The amplified transmission signal outputs are sent to antennas 103 via the duplexer/time switch 107.

Narrowband Transponder Structure

Referring to Figure 2, an example of a remote transponder, suitable for use in implementing the present invention is shown. This transponder is designed to be inexpensive and simple. The particular transponder design shown can also be made in a small, portable, and lightweight package that can be used at the installation of the base station, if desired. The transponder can be mounted on a nearby fixture or even on the antenna mast that is used by the base station's antennas. Alternatively, the transponder can instead be operated as a special mode of a much more complex and fully functional user terminal. A second base station can also perform the transponder functions. The function of the transponder 118 is to receive a signal in the range of the wideband downlink channel, up-convert or down-convert it to the wideband uplink channel, filter it to select only a narrow frequency band, amplify it, and then re-transmit it as a signal in the range of the uplink channel. As mentioned above, frequency-shifting transponder 118 is only one possible example of a transponder suitable for use in calibration. The only general requirement for the transponder is that it transmits back a radio frequency signal that is somehow distinguishable from the signal it received. Besides frequency shifting the signal, the transponder can also time delay the signal, or more generally modulate it with various well-known modulation schemes. For a code division multiplex system, the transponder can also decode the received signal and encode it with a new spreading code for the uplink channel.

As shown in Figure 2, the calibration signal from the base station is received at the transponder antenna 122. A duplexer 140 separately routes signals received at the antenna to the receive chain beginning with a receive bandpass filter 126 and signals coming from the transmit chain, ending with a transmit bandpass filter 125. In the receive chain, signals coming from the transponder antenna after filtering 125 are routed to a low noise amplifier (LNA) 142. This amplified signal is then filtered again by a bandpass filter 144, which eliminates unwanted signals based on their frequencies. This filtered signal is then down-converted to IF (intermediate frequency) by a mixer 148 that combines the received signal with a LO (local oscillator signal) 146 waveform. The IF signal is processed through another bandpass filter 150 before upconversion for transmission. The channel filter 150 can be configured to have two or more passbands, one for each of the frequencies of the calibration signal from the base station.

A second mixer 149 combines the signals from the bandpass filter 150 and a second LO 147 to produce two new transmit signals at frequencies spaced apart from each other and within the uplink frequency band. These two new signals are bandpass filtered 145 and amplified in a power amplifier 143. The power amplifier is adjusted by a power feedback control loop 141 to reduce interference with other channels and smooth reception of the calibration signal at the base station. Another bandpass filter 125 eliminates the upper mixer product and any artifacts from the power amplifier, leaving only the lower mixer product which is a copy of the original input signal on the RF receive chain except for its frequency. This signal is connected to the duplexer 140 for transmission through the antenna element 122. The transponder shows, as an alternative, a separate transmit antenna element 123 and receive antenna element 124. If separate elements are used then the duplexer 140 is no longer required and the antennas can be directly coupled to the respective transmit and receive bandpass filters.

The transponder described above is designed to shift and transpond narrowband signals from the base station that are transmitted in the band for North American cellular CDMA communications, designated as IS-95 by the Telecommunications Industry Association (TIA). In some circumstances, it might be desirable to receive a wideband calibration signal over the complete CDMA channel and return it as a narrowband signal. Since most single channel communication bandwidths are too wide for practical filters at RF frequencies, such a single channel transponder would mix the RF frequency down to a lower intermediate frequency,

apply a narrowband filter at this intermediate frequency, and then mix the filtered signal back up to the desired RF frequency to be echoed back as a narrowband signal. In all other aspects, the wideband, single channel, transponder would behave and be constructed like the narrowband transponder described here.

To determine group delay, at least two frequencies of the calibration signal are desired. To return the two frequencies of the calibration signal, the transponder can be configured to return the two narrowband signals shifted in frequency. Alternatively an additional transponder with unique or some shared hardware can be used. Each transponder can be configured to receive and transmit only in a narrow band or to receive and transmit a broad range of different frequencies. The particular design of the multiple frequency transponder system will depend on the particular circumstances of the application and the communication system.

In operation, the base station DSP 217 generates a specialized narrowband calibration transmit signal on at least two frequencies which it transmits from the antenna array through the duplexer. The transponder receives the calibration transmit signal and echoes it back with the appropriate changes so that it will be received through the receive chain through the duplexer. In a conventional cellular CDMA system, the radio system uses different frequencies for transmit and receive. Thus, the transponder echoes back a signal on the uplink frequency band that is a frequency-shifted copy of the downlink signal it receives. The base station DSP acquires the echoed calibration signal on both frequencies through the receive chain and uses this received calibration signal along with knowledge of the transmit calibration signal to calculate group delay vectors which are then stored in a group delay calibration vector storage buffer.

For a CDMA cellular system, the system may be allocated a bandwidth from, e.g., 824 MHz to 835 MHz or from 835MHz to 849 MHz. The wideband channels within this range may be as narrow as 1.25 MHz or as wide as 5 MHz. In such a system, uplink and downlink frequency bands are typically separated from each other with a significant guard band so that they are separated by 1.25 MHz to 5 MHz. This is the amount by which the transponder must shift the calibration signal frequency to send it back to the base station. In other systems, the wideband uplink and downlink channels may be as wide as 40 MHz or more. The narrowband calibration signals on the other hand, would typically be from 0.01 MHz to 0.1 MHz wide. The spectral width of the calibration signal will be as small as reasonably convenient with readily available equipment at moderate cost. The narrower the signal, the less it will

interfere with existing traffic. However, as mentioned above, the narrowband signal must also be able to be transmitted and received by the wideband transmit and receive chains. The necessary bandwidth limitations will also depend on the particular system. For a system in which the wideband signals are 1.25 MHz wide, the narrowband signals will probably be much narrower than for a system in which the wideband signals are 40 MHz wide. The particular carrier frequencies used can also be adapted to suit the needs of the particular system. Currently, appropriate systems have carrier frequencies centered at frequencies ranging from 450 MHz to 2100 MHz. This range is expected to become greater as radio technologies and spectrum allocations change.

Calculation of Calibration Vectors

There are a variety of different ways to calculate and calibrate the phases and amplitudes of a multiple antenna array using narrowband signals and a transponder. U.S. Patents Nos. 5,546,090 issued August 13, 1996 to Roy, III et al., 5,930,243 issued July 27, 1999 to Parish et al. and 6,037,898 issued to Parish et al. show suitable approaches to calibration. Another approach is shown in International Application No. WO99157820, published November 11, 1999 of Boros et al. The disclosures of these references are hereby incorporated by reference herein.

With respect to calibrating the group delay for the transmit and receive chains of the base station, assuming identical RF propagation on the uplink and downlink, a single transponder or subscriber unit can be used together with its base station to carry out the calibration. However, the present invention enables the separate determination of the uplink and downlink signatures for the transponder or any subscriber unit. These spatial signatures include the effects of the electronic signal paths in the base station hardware and any differences between the uplink and downlink electronic signal paths for the transponder or subscriber unit. One use of such information is to determine separate calibrations for each subscriber unit when the RF propagation to and from the subscriber unit is different. Another use is for calibrating the base station, but rather than obtaining a single calibration vector using the base station and a single transponder, using several transponders to determine the single calibration vector.

In one embodiment, the single calibration vector is the average calibration vector. In another embodiment, it is the weighted average calibration vector. The weighting given to the estimate made using a particular subscriber unit will depend on

a measure of the quality of the signal received by that subscriber unit, so that estimates from subscriber units having better quality signals are weighed more in the weighted average. A method and apparatus for determining signal quality is disclosed in International Application No. WO99/40689, published August 12, 1999 of Yun.

In the architecture of Figures 1 and 2, the base station DSP generates a set of signals that are used for calibration. In one example, all antennas transmit different known calibration signals so that the channel from each transmit antenna to each receive antenna can be calculated. Generally, after subtracting out the components specific to the transponder's location, a receive calibration vector can then be estimated from the difference in phase and amplitude with frequency of the channels from one transmit antenna to each receive antenna. By averaging the results from all the transmit antennas, the calibration vector can be improved still further. Correspondingly, a calibration vector of the transmit chains can be estimated, after subtracting out the transponder specific components, from the relative phases and amplitudes of the channels from different transmit antennas to one of the receive antennas. Again, averaging the results from all the different receive antennas can improve the estimate.

Using the two or more narrow band transponder returns, the relative phase and amplitude of the transmit and receive chains can be calibrated at two frequencies within the base station downlink and uplink bands, respectively. The measurements can also be used for calibrating group delay and any other frequency dependent differences between the receive or transmit chains. Higher accuracy can be obtained if the two narrow frequency bands are placed some distance apart within the traffic bands. Higher accuracy can also be obtained by using more than two different frequencies. The best choice of calibration frequencies and numbers of different frequencies will depend on the bandwidth of the traffic bands and the desired accuracy.

Because a group delay can be regarded as equivalent to a phase ramp with a specific slope, the relative difference in group delay among the transmit and receive chains, respectively, can be calibrated using the phase measurements. This can be done by computing the slopes of the phase ramps based on the phase measurements at the two frequencies within the bands. Since there is an ambiguity in each phase measurement due to phase wrapping, the relative phase between the two measurement frequencies can only be determined to within a phase window of 360 degrees. As a result, any group delay changes and differences within the delay corresponding to a

phase shift of 360 degrees between the two measurement frequencies can be measured and compensated for.

The group delay can be determined directly from a phase calibration process. If the system is calibrating the various receive and transmit chains for phase and amplitude differences, the phase determinations from that process can be used to find the group delay. Group delay can also be determined using relative phase measurements that are calculated apart from any phase calibration process. The phase calibration will give a calibration vector with a calibration coefficient α_{ij} for each antenna i and frequency j. The actual phase φ_{ij} of an antenna i at frequency j can be expressed as $\varphi_{ij} = \alpha_{ij} + \delta_j$, where δ_j is an arbitrary unknown phase term that is common to all antennas at frequency j. The value of δ need not be known in order to calibrate the transmit or receive chain with respect to the other chains. Only the relative phases characterized by the α's is needed.

For group delay, the difference between different transmit or receive chains is used. For a single frequency j, this difference $\Delta\varphi_j$ between antenna i and i' can be expressed as $\Delta\varphi_j = \varphi_{ij} - \varphi_{i'j} = \alpha_{ij} + \delta_j - (\alpha_{i'j} + \delta_j) = \alpha_{ij} - \alpha_{i'j}$. The group delay between the antennas i and i' is obtained by comparing the difference in phase $\Delta\varphi$ at different frequencies. For frequencies j and j', the group delay is therefore proportional to $\Delta\varphi_j - \Delta\varphi_{j'}$. Using the phase calibration vectors α's at the two different frequencies, the relative group delay can quickly be determined.

In the process described above, δ_j the arbitrary unknown phase term that is common to all antennas at frequency j remains unknown. This term can also vary over time. For example if frequency f_1 is repeatedly measured, the measured signature can be expressed as $e^{j\varphi}\mathbf{a}_1$, where $\mathbf{a}$ is the measurement vector at frequency f_1 containing elements a_1, a_2, a_3, ... and the phase φ changes with each measurement. Alternatively, the measured phase can be normalized so that some component, for example, the first component, is real. in either case, the absolute phase is not measured.

As a result, the absolute group delay cannot easily be determined using the phase calibration values, however correcting for relative phase delays between the different transmit and receive chains significantly enhances performance. These relative phase differences constitute the differential phase delay between the transmit and receive chains of the system. Current digital signal processing technology can accommodate a frequency dependent phase variation from a single transmitter. If the phase variations from multiple transmitters can be aligned, then the variations in the

multiple transmitter system can be accommodated by the receiver in the same way as from a single transmitter. If the phase variations differ among the transmitters, the transmitted signal becomes much more difficult to resolve. Accordingly while a calibration that corrects for absolute group delay may be desirable in some applications, calibration for relative group delay is very useful. The more the differences between the transmit or alternatively, receive chains, can be reduced the higher the system's performance.

Using phase and amplitude measurements, calibration vectors can be formed and applied to transmissions by the base station. One approach uses spatial signatures from the receive chains of an antenna system and, using signatures at two different frequencies imposes a linear phase shift ramp. The spatial signatures can be made up of a vector or a set **a** of phase and amplitude measurements for each receive or transmit chain. They can be represented as $\mathbf{a}_j$ and $\mathbf{a}_{j'}$, where $\mathbf{a}_j$, for example, represents a set of values a_{j1}, a_{j2}, a_{j3}, ... a_M for each of M receive or transmit chains i = 1, 2, 3, ... M, at the frequency j. These two signatures are combined to derive the frequency dependent calibration factor $\mathbf{c}(f)$.

While a linear fit for $\mathbf{c}(f)$ provides for a simple and quick determination of the calibration vector using only two measured frequencies, as shown below, more frequencies can be measured and any variety of other curves or shapes can be matched to the measured results. The choice of an interpolation or curve matching algorithm as well as the choice of the number of different frequencies to measure will depend on a balance between calibration complexity and signal quality. The quality of the equalizers and the demodulators as well as the width of the frequency bandwidth of the system will likely also be considered among other factors.

To calibrate differential amplitude shifts with frequency, a frequency dependent amplitude calibration factor $| g_i(f) |$ for each antenna i = 1, ... M can be determined by linear interpolation:

$$| g_i(f) | = [(f-f_1)/(f_2-f_1)|a_{1,i}|] + [(f_2-f)/(f_2-f_1)|a_{2,i}|]$$

for $f_2 \geq f \geq f_1$, where f_2 corresponds to frequency j', f_1 corresponds to frequency j, $a_{1,i}$ corresponds to the phase and amplitude measurement for antenna i at frequency f_1 and $a_{2,i}$ corresponds to the phase amplitude measurement for antenna i at frequency f_2. Linear extrapolation can be used to extend the amplitude calibration factor outside the interval between the two measured frequencies f_1, f_2.

To determine a phase portion of the calibration vector $\mathbf{c}(f)$, a modified linear interpolation that compensates for the phase wrapping can be used. As mentioned

above, there is a relative phase window of 360 degrees or 2π, at which point, the phase wraps back around to zero. If angle (a) is an angle in degrees that can take any value from -180 degrees up to but not including 180, angle (a) $\in$ (-180, 180], and angle (a) corresponds to the complex number a, and a* is the complex conjugate of a, then the calibration phase $\varphi_i(f)$ for antenna i at frequency f can be expressed as shown below.

$$\varphi_i(f) = [\ (f-f_1)/(f_2-f_1)\ ]\ \text{angle}\ (a_{1,i})^* a_{2,i} + \text{angle}\ (a_{1,1})$$

for i = 1, ... M and the overall calibration factor is equal to the combination of the amplitude and phase calibration factors which can be expressed as shown below:

$$c_i(f) = |\ g_i(f)\ | e^{j(180/\pi)\varphi i(f)}$$

<u>Method of Operation</u>

An example of an operational process for calibrating a group of receive chains for group delay is shown in Figure 3. Other frequency dependent calibration vectors can be determined using a similar process. The calibration process typically includes calibrating the receive chain and the transmit chain with the same set of samples. Calibration of the transmit chains is shown in Figure 4. To begin a calibration cycle for the receive chain, the base station (BS) (see e.g. Figure 1) will generate a calibration signal. As discussed above, this is typically a narrowband signal at two or more frequencies. This narrowband transmit calibration signal is then transmitted from a single transmit chain of the base station 311. The transmission can occur at any time during the regular use of the base station for normal operation due to the small amount of additional energy added to the existing wideband data traffic by the narrowband signal. While only one transmit chain is required, transmitting from all of the transmit chains at once provides more samples for the receive calibration algorithms.

The transmitted narrowband calibration signal is received at the transponder 313, (see e.g. Figure 2). If the calibration signal is a wideband signal, it is converted to a set of at least two narrowband waveforms using appropriate bandpass filters as discussed above. If the signal has a particular spreading sequence or is modulated with a particular data or training sequence, this can be demodulated and a new signal can be modulated onto the signal. In one embodiment, the calibration signal is a narrowband signal, which is simply received, shifted in frequency 315, and transmitted back to the base station 317. This approach simplifies the transponder and

eliminates many other potential causes of errors. The frequency shifted calibration signal can also be shifted to two or more different frequencies and retransmitted so that calibration can be performed across different narrow frequency bands. However, the same effect can be achieved with a simpler transponder by sending several different calibration signals from the base station, each at a different frequency for the downlink. Each signal will be shifted to a different frequency for the uplink.

The base station receives the transponder signal at each of its receive antenna chains 319. These received transponder signals are sampled for each receive antenna chain 321 and the samples can be used to measure any number of characteristics of the received signal. Each set of samples from each receive chain represents a different view of the same narrowband transponder signal. To enhance reception, the DSP 217 will typically use narrow bandpass filters to eliminate most of the data traffic signal energy and isolate the received transponder signal. The received transponder signal is used to calculate a set of phases, for example the α's discussed above and amplitudes 323 The calculation in support of group delay will typically be based on comparing the received transponder signal as it was received by each receive chain to each signal as received by each other receive chain. This is commonly done by measuring phases and amplitudes and using a covariance matrix, for example. As an alternative, the signal can be sampled at only two receive chains. This will allow the two selected chains to be calibrated against each other. By repeating the process for each possible combination or for each receive chain against a receive chain selected to be the reference, a set of relative phase measurements can be obtained.

The process of transmitting and receiving calibration signals described above can then be repeated and the results averaged or stored 325. Further relative phases and amplitudes are calculated using the additional data 327 and a group delay is calculated 328. This group delay is typically in the form of a calibration vector composed of a set of phase and amplitude correction factors for each transmit and receive chain, as discussed above. Alternatively, the resulting calibration vector can be applied and the process repeated to find a new vector that is used to adjust the first vector. By applying the adjusted calibration vector after each cycle, the calibration should become progressively more accurate until it converges on the limit of the calibration system's accuracy. The transmission, reception and computations can be repeated for different combinations of receive chains and even for different transponders. Over time, the characteristics of the receive chains can change and so the process can also be repeated in order to update the calibration vectors with

changing conditions. When the calibrations are done against a reference chain, pairing each receive chain against the reference, the reference chain's vectors can be set at one, or some other normalized set of values, so that the vectors for the other receive chains represent the variance from the reference chain. Alternatively, the vectors can represent the variance from any other value, for example an average, mean or median response.

Calibration of the transmit chain is done in a similar way as shown in Figure 4. As with the receive chain, a calibration signal is transmitted to the transponder. In this case, the calibration signal is transmitted from each of the base station's transmit chains 329. So that they can be distinguished from each other when received, each receive chain uses a different modulation sequence. As with the receive calibration, this signal is a narrowband signal at at least two different frequencies. The narrowband signal allows the transponder to have a simple construction.

The calibration signals are received at the transponder 331. Which then, as with the receive calibration, shifts the frequency of the received calibration signals 333. After that, the shifted calibration signals are transmitted back to the base station 335. It is again possible to change modulated sequences or spreading codes but the simplest transponder will take the narrowband signal that it receives in the downlink band and transmit it back as a virtually identical narrowband signal in the uplink band.

The base station receives the transponder signals this time at just one receive antenna chain 337. The received transponder signals are sampled 339 and then the unique modulated sequences are used to extract each transmit chain calibration signal 341 from the sampled waveform. As with the receive calibration, a narrow bandpass filter is typically used to isolate the transponder signal. For calibration purposes, the transmitted calibration signals from each transmit chain are compared to each other 343. In order to make it easier to distinguish the simultaneously received signals from the different transmit chains, the number of simultaneous transmit chains can be reduced. For example, one of the transmit chains can be designated as the reference and then each other transmit chain can transmit with the reference, one pair at a time, until all the transmit chains have been calibrated against the reference. This is similar to the pair-wise receive chain calibration mentioned above.

These comparisons become the basis for generating a set of relative phases and amplitudes 345. The process of sending and receiving calibration signals can then be repeated 347 and further relative phases and amplitudes computed 349 to

refine the results. Then, the transmit group delay calibration vector can be calculated for each transmit chain 351. In one embodiment, the calibration vector determined in the first round is applied to each transmit chain, and then the process is repeated. The next calibration cycle will lead to greater accuracy as the gross errors have already been compensated. This is similar to performing a coarse tuning process and then a fine-tuning process.

The present invention provides many advantages over the prior art. Calibrations can be performed using only a simple, inexpensive transponder. Both transmit and receive calibration can be determined in a single transaction and the method self-corrects for reference frequency offsets in the antenna array system. Accordingly, calibration in accordance with the present invention is inherently accurate. While the invention has been described primarily as a calibration of a base station using a remote transponder, it can be applied to remote user terminals that have multiple antennas. It can also be applied to any other type of wireless network with multiple antenna system whether one with base stations and remotes, equal peers or masters and slaves.

To improve the reception of regular traffic during calibration, it may be desirable to apply a notch filter at the base station to filter out the transponder signal bands. This would typically be a digital filter and can be turned off when no calibration signal is active. The subscriber units could similarly have a notch filter for the calibration signal from the base station.

In the description above, for the purposes of explanation, numerous specific details are set forth in order to provide a thorough understanding of the present invention. It will be apparent, however, to one skilled in the art that the present invention may be practiced without some of these specific details. In other instances, well-known structures and devices are shown in block diagram form.

The present invention includes various steps. The steps of the present invention may be performed by hardware components, such as those shown in Figures 1 and 2, or may be embodied in machine-executable instructions, which may be used to cause a general-purpose or special-purpose processor or logic circuits, such as a DSP programmed with the instructions to perform the steps. Alternatively, the steps may be performed by a combination of hardware and software.

The present invention may be provided as a computer program product which may include a machine-readable medium having stored thereon instructions which may be used to program a computer (or other electronic devices) to perform a process

according to the present invention. The machine-readable medium may include, but is not limited to, floppy diskettes, optical disks, CD-ROMs, and magneto-optical disks, ROMs, RAMs, EPROMs, EEPROMs, magnet or optical cards, flash memory, or other type of media or machine-readable medium suitable for storing electronic instructions. Moreover, the present invention may also be downloaded as a computer program product, wherein the program may be transferred from a remote computer to a requesting computer by way of data signals embodied in a carrier wave or other propagation medium via a communication link (e.g., a modem or network connection).

Importantly, while the present invention has been described in the context of a wireless spread spectrum data system for mobile remote terminals, it can be applied to a wide variety of different wireless systems in which data is exchanged. Such systems include voice, video, music, broadcast and other types of data systems without external connections. The present invention can be applied to fixed user terminals as well as to low and high mobility terminals. Many of the methods are described herein in a basic form but steps can be added to or deleted from any of the methods and information can be added or subtracted from any of the described messages without departing from the basic scope of the present invention. It will be apparent to those skilled in the art that many further modifications and adaptations can be made. The particular embodiments are not provided to limit the invention but to illustrate it. The scope of the present invention is not to be determined by the specific examples provided above but only by the claims below.

CLAIMS

What is claimed is:

1. A radio communications system comprising:

an antenna array adapted to transmit and receive radio communications signals with a plurality of other terminals, the communications signals each using a particular minimum bandwidth;

a transmit chain to transmit a calibration signal through the antenna array to a transponder on at least two different frequency bands within the minimum bandwidth;

a receive chain to receive through the antenna array a transponder signal from the transponder, the transponder signal being received on at least two different frequency bands and being based on the calibration signal; and

a signal processor to determine a frequency dependent calibration vector based on the at least two frequency bands of the transponder signal as received through the receive chain.

2. The system of claim 1, wherein determining a frequency dependent calibration vector comprises comparing relative phases for the transponder signal at a first one of the at least two frequencies to relative phases for the transponder signal at a second one of the at least two frequencies to determine a group delay.

3. The system of claim 1, wherein the transponder signal is shifted in frequency as compared to the calibration signal.

4. The system of claim 1, further comprising measuring the relative phases and amplitudes of the transponder signal as received by the receive chain.

5. The system of claim 4:

wherein the receive chain comprises a plurality of receive chains;

wherein each receive chain receives the transponder signal; and

wherein the signal processor determines a group delay by comparing the relative phases of the transponder signal at each frequency as received by each receive chain.

6. The system of claim 5 wherein determining a frequency dependent calibration vector comprises determining a receive chain group delay by comparing a phase difference between at least two receive chains for the transponder signal at a first one of the at least two frequency bands to a phase difference between the same two receive chains for the transponder signal at a second one of the at least two frequency bands.

7. The system of claim 6, wherein one of the plurality of receive chains is selected as a reference receive chain and the group delay for each receive chain is characterized with respect to the reference receive chain.

8. The system of Claim 4 wherein the signal processor determines an uplink signature of the transponder at the antenna array at each frequency of the transponder signal using measured phases and amplitudes of the transponder signal and wherein the signal processor determines the frequency dependent calibration vector for the receive chain using the uplink signatures of the transponder.

9. The system of Claim 4 wherein the signal processor determines a downlink signature of the transmit chain at the transponder using measured phases and amplitudes at each frequency of the transponder signal and wherein the signal processor further determines the frequency dependent calibration vector for the transmit chain using the downlink signatures of the transmit chain.

10. The system of claim 1:

wherein the transmit chain comprises a plurality of transmit chains;

wherein each transmit chain transmits the calibration signal; and

wherein the signal processor determines a frequency dependent transmit calibration vector by comparing the relative phases of the transponder signal at each frequency of the transponder signal as received by each receive chain.

11. The system of claim 10, wherein the calibration signal comprises a plurality of signals, one from each transmit chain, each signal being individually identifiable based on a unique modulation sequence.

12. The system of claim 10 wherein determining a frequency dependent transmit calibration vector comprises comparing a phase difference between two transmit chains for the transponder signal at a first one of the at least two frequencies to a phase difference between the same two transmit chains for the transponder signal at a second one of the at least two frequencies to determine a group delay.

13. The system of Claim 12 wherein one of the plurality of transmit chains is selected as a reference chain and the group delay of each transmit chain is defined with respect to the reference chain.

14. A machine-readable medium having stored thereon data representing instructions which, when executed by a machine, cause the machine to perform operations comprising:

transmitting radio communications signals to a plurality of other terminals using a transmit chain, the communications signals each using a particular minimum transmit bandwidth;

receiving radio communications signals from a plurality of other terminals using a receive chain, the communications signals each using a particular minimum receive bandwidth;

transmitting a calibration signal through the transmit chain to a transponder on at least two different frequency bands within the minimum transmit bandwidth;

receiving a transponder signal through the receive chain from the transponder, the transponder signal being received on at least two different frequency bands within the minimum receive bandwidth and being based on the calibration signal; and

determining a frequency dependent calibration vector based on the at least two frequency bands of the transponder signal as received through the receive chain.

15. The medium of claim 14, wherein determining a frequency dependent calibration vector comprises comparing relative phases for the transponder signal at a first one of the at least two frequencies to relative phases for the transponder signal at a second one of the at least two frequencies to determine a group delay.

16. The medium of claim 14, wherein the transponder signal is shifted in frequency as compared to the calibration signal.

17. The medium of claim 14 wherein determining a frequency dependent calibration vector comprises determining a receive chain group delay by comparing a phase difference between at least two receive chains for the transponder signal at a first one of the at least two frequency bands to a phase difference between the same two receive chains for the transponder signal at a second one of the at least two frequency bands.

18. The medium of claim 14 wherein determining a frequency dependent transmit calibration vector comprises comparing a phase difference between two transmit chains for the transponder signal at a first one of the at least two frequencies to a phase difference between the same two transmit chains for the transponder signal at a second one of the at least two frequencies to determine a group delay.

19. A method comprising:

transmitting radio communications signals to a plurality of other terminals using a transmit chain, the communications signals each using a particular minimum transmit bandwidth;

receiving radio communications signals from a plurality of other terminals using a receive chain, the communications signals each using a particular minimum receive bandwidth;

transmitting a calibration signal through the transmit chain to a transponder on at least two different frequency bands within the minimum transmit bandwidth;

receiving a transponder signal through the receive chain from the transponder, the transponder signal being received on at least two different frequency bands within the minimum receive bandwidth and being based on the calibration signal; and

determining a frequency dependent calibration vector based on the at least two frequency bands of the transponder signal as received through the receive chain.

20. The method of claim 19, wherein determining a frequency dependent calibration vector comprises measuring the relative phases and amplitudes of the transponder signal as received by the receive chain.

21. The method of claim 19, wherein determining a frequency dependent calibration vector comprises determining a group delay by comparing the relative phases of the transponder signal at each frequency as received by a plurality of receive chains.

22. The method of claim 21, wherein one of the plurality of receive chains is selected as a reference receive chain and the group delay for each receive chain is characterized with respect to the reference receive chain.

23. The method of Claim 21 wherein determining a frequency dependent calibration vector comprises determining an uplink signature of the transponder at the receive chains at each frequency of the transponder signal using measured phases and amplitudes of the transponder signal and determining the frequency dependent calibration vector for the receive chains using the uplink signatures of the transponder.

24. The method of Claim 21 wherein the signal processor determining a frequency dependent calibration vector comprises determining a downlink signature of a plurality of transmit chains at the transponder using measured phases and amplitudes at each frequency of the transponder signal and determining the frequency dependent calibration vector for the transmit chains using the downlink signatures of the transmit chains.

25. The method of claim 19 wherein determining a frequency dependent calibration vector comprises determining a frequency dependent transmit calibration

vector by comparing the relative phases of the transponder signal at each frequency of the transponder signal as received by each of a plurality of receive chains.

26. The system of claim 1, wherein the system is a code division multiple access system.

27. The medium of claim 14, wherein the radio communications signals conform to a standard for code division multiple access.

28. The method of claim 19, wherein the radio communications signals conform to a standard for code division multiple access.

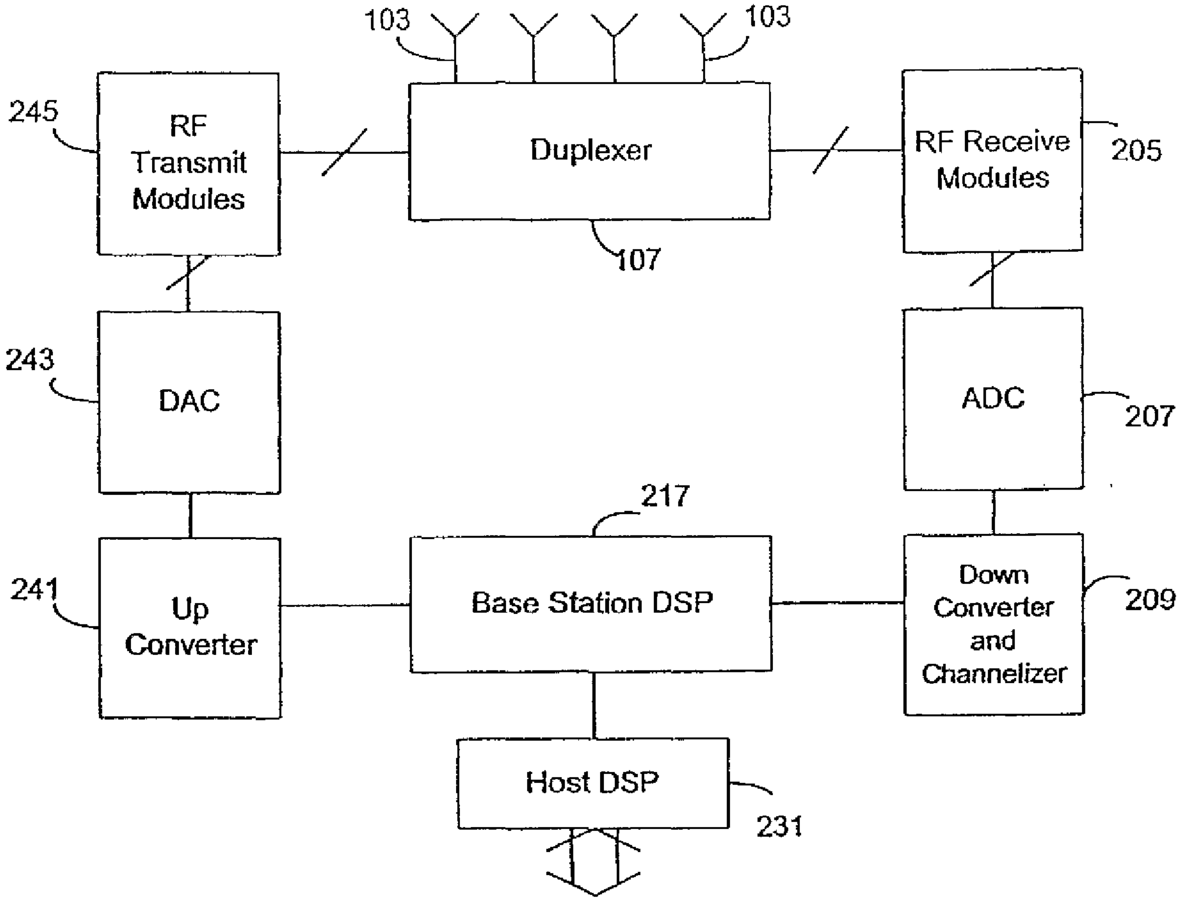
103
103
245
RF Transmit Modules
Duplexer
RF Receive Modules
205
107
243
DAC
ADC
207
217
241
Up Converter
Base Station DSP
Down Converter and Channelizer
209
Host DSP
231

Figure 1

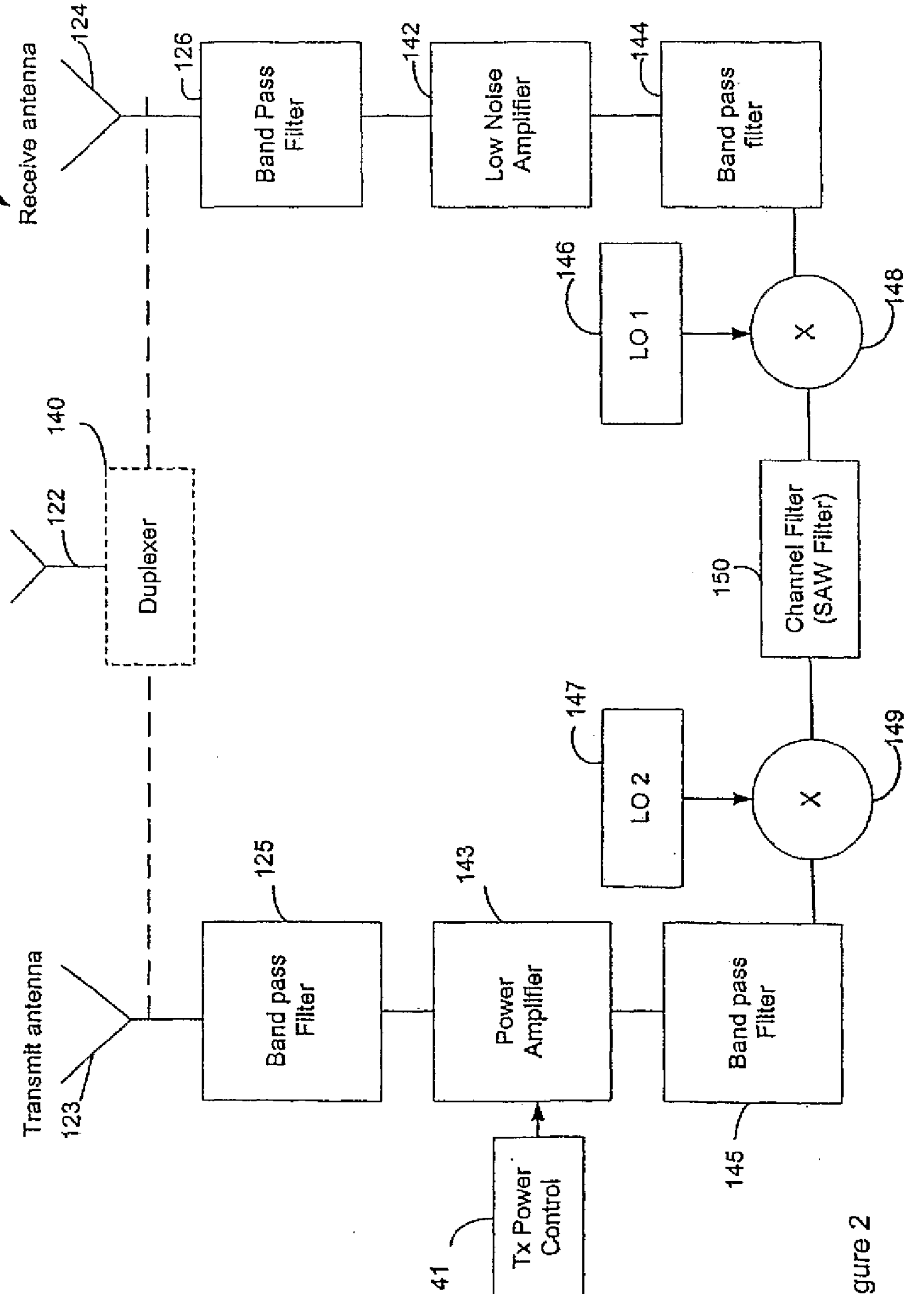
118
Receive antenna
124
126
Band Pass Filter
142
Low Noise Amplifier
144
Band pass filter
146
LO 1
148
140
Duplexer
122
150
Channel Filter (SAW Filter)
147
LO 2
149
Transmit antenna
123
125
Band pass Filter
143
Power Amplifier
Band pass Filter
145
141
Tx Power Control

Figure 2

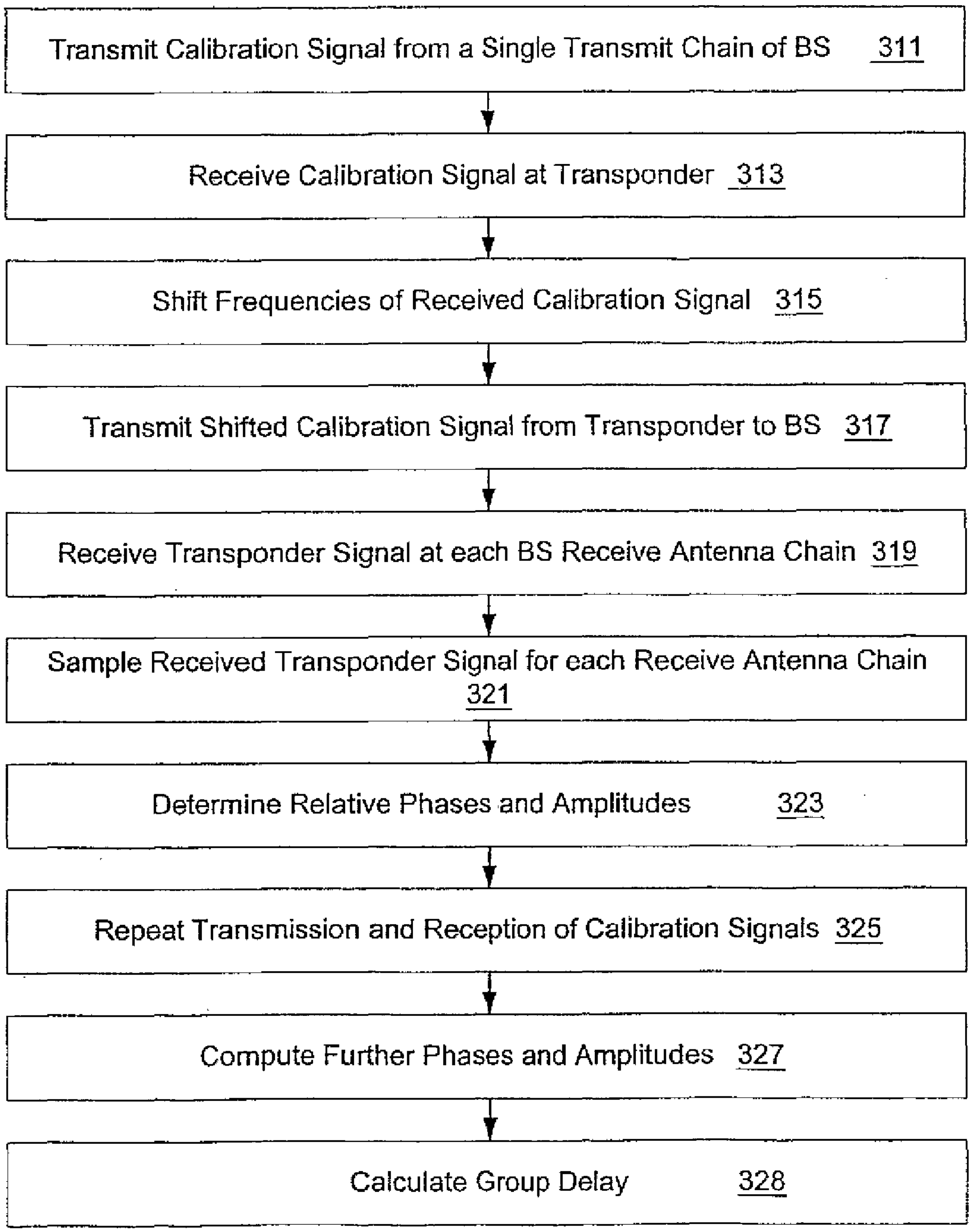
Transmit Calibration Signal from a Single Transmit Chain of BS 311
Receive Calibration Signal at Transponder 313
Shift Frequencies of Received Calibration Signal 315
Transmit Shifted Calibration Signal from Transponder to BS 317
Receive Transponder Signal at each BS Receive Antenna Chain 319
Sample Received Transponder Signal for each Receive Antenna Chain 321
Determine Relative Phases and Amplitudes 323
Repeat Transmission and Reception of Calibration Signals 325
Compute Further Phases and Amplitudes 327
Calculate Group Delay 328

Figure 3

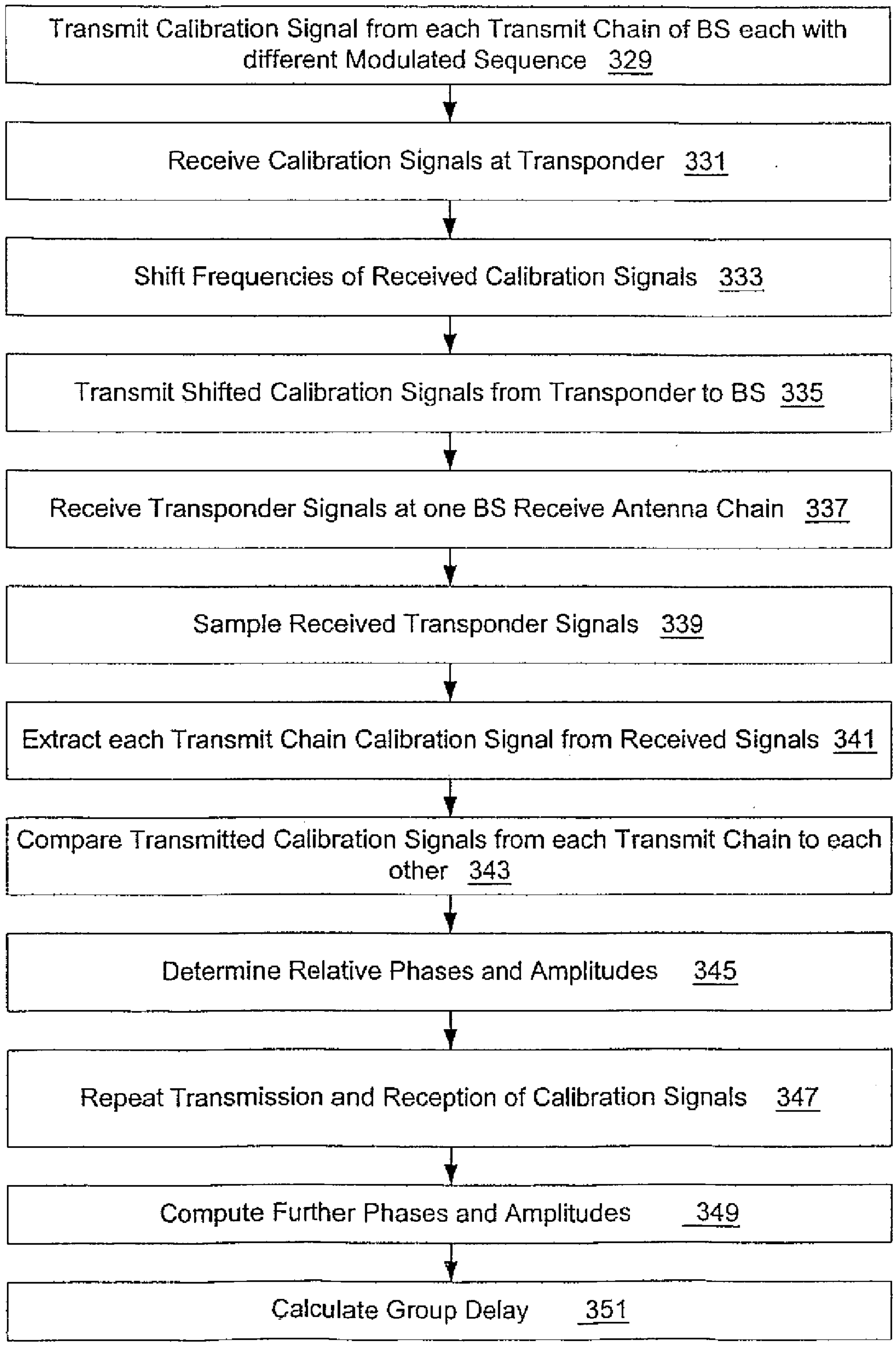
Transmit Calibration Signal from each Transmit Chain of BS each with different Modulated Sequence 329
Receive Calibration Signals at Transponder 331
Shift Frequencies of Received Calibration Signals 333
Transmit Shifted Calibration Signals from Transponder to BS 335
Receive Transponder Signals at one BS Receive Antenna Chain 337
Sample Received Transponder Signals 339
Extract each Transmit Chain Calibration Signal from Received Signals 341
Compare Transmitted Calibration Signals from each Transmit Chain to each other 343
Determine Relative Phases and Amplitudes 345
Repeat Transmission and Reception of Calibration Signals 347
Compute Further Phases and Amplitudes 349
Calculate Group Delay 351

Figure 4

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 02/30896

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H01Q3/26

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H01Q

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, WPI Data, INSPEC, COMPENDEX

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	US 5 546 090 A (ROY III RICHARD H ET AL) 13 August 1996 (1996-08-13) cited in the application abstract column 2, line 13-39 column 4, line 20 -column 8, line 53 figures 4-7 ---	1,3,4, 14,16, 19,20, 26-28
X	US 5 294 934 A (MATSUMOTO SOICHI) 15 March 1994 (1994-03-15) abstract column 4, line 63 -column 6, line 7 figure 1 --- -/--	1,14,19

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
11 December 2002	20/12/2002

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31–70) 340–2040, Tx. 31 651 epo nl, Fax: (+31–70) 340–3016	Helms, J

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 02/30896

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
X	US 6 236 839 B1 (GU YUCONG ET AL) 22 May 2001 (2001-05-22) abstract column 1, line 37 -column 4, line 8 figure 1	1,14,19
A	US 6 037 898 A (BARRATT CRAIG H ET AL) 14 March 2000 (2000-03-14) cited in the application abstract column 9, line 49 -column 18, line 65 figure 2	1-28

INTERNATIONAL SEARCH REPORT

Information on patent family members

International Application No
PCT/US 02/30896

Patent document cited in search report		Publication date	Patent family member(s)			Publication date
US 5546090	A	13-08-1996	US	5515378	A	07-05-1996
			WO	9818272	A1	30-04-1998
			US	5592490	A	07-01-1997
			US	5828658	A	27-10-1998
			AT	183865	T	15-09-1999
			AU	670766	B2	01-08-1996
			AU	3145493	A	19-07-1993
			CA	2125571	A1	24-06-1993
			DE	69229876	D1	30-09-1999
			DE	69229876	T2	30-12-1999
			EP	0616742	A1	28-09-1994
			EP	0926916	A2	30-06-1999
			ES	2137197	T3	16-12-1999
			FI	942771	A	10-06-1994
			JP	7505017	T	01-06-1995
			JP	3266258	B2	18-03-2002
			WO	9312590	A1	24-06-1993
			US	5625880	A	29-04-1997
			US	5642353	A	24-06-1997
US 5294934	A	15-03-1994	JP	5136622	A	01-06-1993
			FR	2684505	A1	04-06-1993
			GB	2262386	A ,B	16-06-1993
US 6236839	B1	22-05-2001	WO	0119101	A1	15-03-2001
US 6037898	A	14-03-2000	WO	0031892	A1	02-06-2000
			AU	2198099	A	13-06-2000
			EP	1133836	A1	19-09-2001
			JP	2002530998	T	17-09-2002

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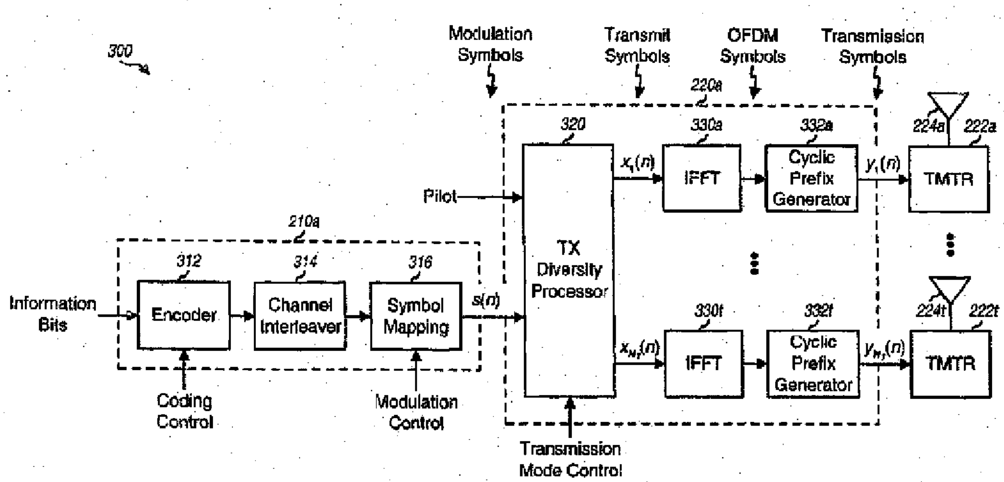
(54) Title: DIVERSITY TRANSMISSION MODES FOR MIMO OFDM COMMUNICATION SYSTEMS



(57) Abstract: Techniques for transmitting data using a number of diversity transmission modes to improve reliability. At a transmitter, for each of one or more data streams, a particular diversity transmission mode is selected for use from among a number of possible transmission modes. These transmission modes may include a frequency diversity transmission mode, a Walsh diversity transmission mode, a space time transmit diversity (STTD) transmission mode, and a Walsh-STTD transmission mode. Each diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof. Each data stream is coded and modulated to provide modulation symbols, which are further processed based on the selected diversity transmission mode to provide transmit symbols. For OFDM, the transmit symbols for all data streams are further OFDM modulated to provide a stream of transmission symbols for each transmit antenna used for data transmission.

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DIVERSITY TRANSMISSION MODES FOR MIMO OFDM COMMUNICATION SYSTEMS

BACKGROUND

Field

[1001] The present invention relates generally to data communication, and more specifically to techniques for transmitting data using a number of diversity transmission modes in MIMO OFDM systems.

Background

[1002] Wireless communication systems are widely deployed to provide various types of communication such as voice, packet data, and so on. These systems may be multiple-access systems capable of supporting communication with multiple users either sequentially or simultaneously. This is achieved by sharing the available system resources, which may be quantified by the total available operating bandwidth and transmit power.

[1003] A multiple-access system may include a number of access points (or base stations) that communicate with a number of user terminals. Each access point may be equipped with one or multiple antennas for transmitting and receiving data. Similarly, each terminal may be equipped with one or multiple antennas.

[1004] The transmission between a given access point and a given terminal may be characterized by the number of antennas used for data transmission and reception. In particular, the access point and terminal pair may be viewed as (1) a multiple-input multiple-output (MIMO) system if multiple (N_T) transmit antennas and multiple (N_R) receive antennas are employed for data transmission, (2) a multiple-input single-output (MISO) system if multiple transmit antennas and a single receive antenna are employed, (3) a single-input multiple-output (SIMO) system if a single transmit antenna and multiple receive antennas are employed, or (4) a single-input single-output (SISO) system if a single transmit antenna and a single receive antenna are employed.

[1005] For a MIMO system, a MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_S independent channels, with $N_S \leq \min\{N_T, N_R\}$. Each of the N_S independent channels is also referred to as a spatial

subchannel of the MIMO channel and corresponds to a dimension. The MIMO system can provide improved performance (e.g., increased transmission capacity and/or greater reliability) if the additional dimensionalities created by the multiple transmit and receive antennas are utilized. For a MISO system, only one spatial subchannel is available for data transmission. However, the multiple transmit antennas may be used to transmit data in a manner to improve the likelihood of correct reception by the receiver.

[1006] The spatial subchannels of a wideband system may encounter different channel conditions due to various factors such as fading and multipath. Each spatial subchannel may thus experience frequency selective fading, which is characterized by different channel gains at different frequencies of the overall system bandwidth. It is well known that frequency selective fading causes inter-symbol interference (ISI), which is a phenomenon whereby each symbol in a received signal acts as distortion to subsequent symbols in the received signal. The ISI distortion degrades performance by impacting the ability to correctly detect the received symbols.

[1007] To combat frequency selective fading, orthogonal frequency division multiplexing (OFDM) may be used to effectively partition the overall system bandwidth into a number of (N_F) subbands, which may also be referred to as OFDM subbands, frequency bins, or frequency sub-channels. Each subband is associated with a respective subcarrier upon which data may be modulated. For each time interval that may be dependent on the bandwidth of one subband, a modulation symbol may be transmitted on each of the N_F subbands.

[1008] For a multiple-access system, a given access point may communicate with terminals having different number of antennas at different times. Moreover, the characteristics of the communication channels between the access point and the terminals typically vary from terminal to terminal and may further vary over time, especially for mobile terminals. Different transmission schemes may then be needed for different terminals depending on their capabilities and requirements.

[1009] There is therefore a need in the art for techniques for transmitting data using a number of diversity transmission modes depending on the capability of the receiver device and the channel conditions.

SUMMARY

[1010] Techniques are provided herein for transmitting data in a manner to improve the reliability of data transmission. A MIMO OFDM system may be designed to support a number of modes of operation for data transmission. These transmission modes may include diversity transmission modes, which may be used to achieve higher reliability for certain data transmission (e.g., for overhead channels, poor channel conditions, and so on). The diversity transmission modes attempt to achieve transmit diversity by establishing orthogonality among multiple signals transmitted from multiple transmit antennas. Orthogonality among the transmitted signals may be attained in frequency, time, space, or any combination thereof. The transmission modes may also include spatial multiplexing transmission modes and beam steering transmission modes, which may be used to achieve higher bit rates under certain favorable channel conditions.

[1011] In an embodiment, a method is provided for processing data for transmission in a wireless (e.g., MIMO OFDM) communication system. In accordance with the method, a particular diversity transmission mode to use for each of one or more data streams is selected from among a number of possible transmission modes. Each diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof. Each data stream is coded and modulated based on coding and modulation schemes selected for the data stream to provide modulation symbols. The modulation symbols for each data stream are further processed based on the selected diversity transmission mode to provide transmit symbols. For OFDM, the transmit symbols for all data streams are further OFDM modulated to provide a stream of transmission symbols for each of one or more transmit antennas used for data transmission. Pilot symbols may also be multiplexed with the modulation symbols using frequency division multiplexing (FDM), time division multiplexing (TDM), code division multiplexing (CDM), or any combination thereof.

[1012] The transmission modes may include, for example, (1) a frequency diversity transmission mode that redundantly transmits modulation symbols over multiple OFDM subbands, (2) a Walsh diversity transmission mode that transmits each modulation symbol over N_T OFDM symbol periods, where N_T is the number of transmit antennas used for data transmission, (3) a space time transmit diversity (STTD) transmission

mode that transmits modulation symbols over multiple OFDM symbol periods and multiple transmit antennas, and (4) a Walsh-STTD transmission mode that transmits modulation symbols using a combination of Walsh diversity and STTD. For the Walsh diversity and Walsh-STTD transmission modes, the same modulation symbols may be redundantly transmitted over all transmit antennas or different modulation symbols may be transmitted over different transmit antennas.

[1013] Each data stream may be for an overhead channel or targeted for a specific receiver device. The data rate for each user-specific data stream may be adjusted based on the transmission capability of the receiver device. The transmit symbols for each data stream are transmitted on a respective group of one or more subbands.

[1014] In another embodiment, a method is provided for processing a data transmission at a receiver of a wireless communication system. In accordance with the method, the particular diversity transmission mode used for each of one or more data streams to be recovered is initially determined. The diversity transmission mode used for each is selected from among a number of possible transmission modes. Received symbols for each data stream are then processed based on the diversity transmission mode used for the data stream to provide recovered symbols, which are estimates of modulation symbols transmitted from a transmitter for the data stream. The recovered symbols for each data stream are further demodulated and decoded to provide decoded data for the data stream.

[1015] Various aspects and embodiments of the invention are described in further detail below. The invention further provides methods, transmitter units, receiver units, terminals, access points, systems, and other apparatuses and elements that implement various aspects, embodiments, and features of the invention, as described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

[1016] The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

[1017] FIG. 1 is a diagram of a multiple-access system that supports a number of users;

[1018] FIG. 2 is a block diagram of an embodiment of an access point and two terminals;

[1019] FIG. 3 is a block diagram of a transmitter unit;

[1020] FIG. 4 is a block diagram of a TX diversity processor that may be used to implement the frequency diversity scheme;

[1021] FIG. 5 is a block diagram of a TX diversity processor that may be used to implement the Walsh diversity scheme;

[1022] FIG. 6 is a block diagram of a TX diversity processor that may be used to implement the STTD scheme;

[1023] FIG. 7 is a block diagram of a TX diversity processor that may be used to implement a repeated Walsh-STTD scheme;

[1024] FIG. 8 is a block diagram of a TX diversity processor that may be used to implement a non-repeated Walsh-STTD scheme;

[1025] FIG. 9 is a block diagram of a receiver unit;

[1026] FIG. 10 is a block diagram of an RX diversity processor;

[1027] FIG. 11 is a block diagram of an RX antenna processor within the RX diversity processor and which may be used for the Walsh diversity scheme; and

[1028] FIG. 12 is a block diagram of an RX subband processor within the RX antenna processor and which may be used for the repeated and non-repeated Walsh-STTD schemes.

DETAILED DESCRIPTION

[1029] **FIG. 1** is a diagram of a multiple-access system 100 that supports a number of users. System 100 includes one or more access points (AP) 104 that communicate with a number of terminals (T) 106 (only one access point is shown in FIG. 1 for simplicity). An access point may also be referred to as a base station, a UTRAN, or some other terminology. A terminal may also be referred to as a handset, a mobile station, a remote station, user equipment (UE), or some other terminology. Each terminal 106 may concurrently communicate with multiple access points 104 when in soft handoff (if soft handoff is supported by the system).

[1030] In an embodiment, each access point 104 employs multiple antennas and represents (1) the multiple-input (MI) for a downlink transmission from the access point to a terminal and (2) the multiple-output (MO) for an uplink transmission from the terminal to the access point. A set of one or more terminals 106 communicating with a given access point collectively represents the multiple-output for the downlink transmission and the multiple-input for the uplink transmission.

[1031] Each access point 104 can communicate with one or multiple terminals 106, either concurrently or sequentially, via the multiple antennas available at the access point and the one or more antennas available at each terminal. Terminals not in active communication may receive pilots and/or other signaling information from the access point, as shown by the dashed lines for terminals 106e through 106h in FIG. 1.

[1032] For the downlink, the access point employs N_T antennas and each terminal employs 1 or N_R antennas for reception of one or more data streams from the access point. In general, N_R can be different for different multi-antenna terminals and can be any integer. A MIMO channel formed by the N_T transmit antennas and N_R receive antennas may be decomposed into N_S independent channels, with $N_S \leq \min\{N_T, N_R\}$. Each such independent channel is also referred to as a spatial subchannel of the MIMO channel. The terminals concurrently receiving downlink data transmission need not be equipped with equal number of receive antennas.

[1033] For the downlink, the number of receive antennas at a given terminal may be equal to or greater than the number of transmit antennas at the access point (i.e., $N_R \geq N_T$). For such a terminal, the number of spatial subchannels is limited by the number of transmit antennas at the access point. Each multi-antenna terminal communicates with the access point via a respective MIMO channel formed by the access point's N_T transmit antennas and its own N_R receive antennas. However, even if multiple multi-antenna terminals are selected for concurrent downlink data transmission, only N_S spatial subchannels are available regardless of the number of terminals receiving the downlink transmission.

[1034] For the downlink, the number of receive antennas at a given terminal may also be less than the number of transmit antennas at the access point (i.e., $N_R < N_T$). For example, a MISO terminal is equipped with a single receive antenna ($N_R = 1$) for downlink data transmission. The access point may then employ diversity, beam

steering, space division multiple access (SDMA), or some other transmission techniques to communicate simultaneously with one or multiple MISO terminals.

[1035] For the uplink, each terminal may employ a single antenna or multiple antennas for uplink data transmission. Each terminal may also utilize all or only a subset of its available antennas for uplink transmission. At any given moment, the N_T transmit antennas for the uplink are formed by all antennas used by one or more active terminals. The MIMO channel is then formed by the N_T transmit antennas from all active terminals and the access point's N_R receive antennas. The number of spatial subchannels is limited by the number of transmit antennas, which is typically limited by the number of receive antennas at the access point (i.e., $N_S \leq \min\{N_T, N_R\}$).

[1036] **FIG. 2** is a block diagram of an embodiment of access point 104 and two terminals 106. On the downlink, at access point 104, various types of traffic data such as user-specific data from a data source 208, signaling, and so on are provided to a transmit (TX) data processor 210. Processor 210 then formats and encodes the traffic data based on one or more coding schemes to provide coded data. The coded data is then interleaved and further modulated (i.e., symbol mapped) based on one or more modulation schemes to provide modulation symbols (i.e., modulated data). The data rate, coding, interleaving, and symbol mapping may be determined by controls provided by a controller 230 and a scheduler 234. The processing by TX data processor 210 is described in further detail below.

[1037] A transmit processor 220 then receives and processes the modulation symbols and pilot data to provide transmission symbols. The pilot data is typically known data processed in a known manner, if at all. In a specific embodiment, the processing by transmit processor 220 includes (1) processing the modulation symbols based on one or more transmission modes selected for use for data transmission to the terminals to provide transmit symbols and (2) OFDM processing the transmit symbols to provide transmission symbols. The processing by transmit processor 220 is described in further detail below.

[1038] Transmit processor 220 provides N_T streams of transmission symbols to N_T transmitters (TMTR) 222a through 222t, one transmitter for each antenna used for data transmission. Each transmitter 222 converts its transmission symbol stream into one or more analog signals and further conditions (e.g., amplifies, filters, and frequency

upconverts) the analog signals to generate a respective downlink modulated signal suitable for transmission over a wireless communication channel. Each downlink modulated signal is then transmitted via a respective antenna 224 to the terminals.

[1039] At each terminal 106, the downlink modulated signals from multiple transmit antennas of the access point are received by one or multiple antennas 252 available at the terminal. The received signal from each antenna 252 is provided to a respective receiver (RCVR) 254. Each receiver 254 conditions (e.g., filters, amplifies, and frequency downconverts) its received signal and further digitizes the conditioned signal to provide a respective stream of samples.

[1040] A receive processor 260 then receives and processes the streams of samples from all receivers 254 to provide recovered symbols (i.e., demodulated data). In a specific embodiment, the processing by receive processor 260 includes (1) OFDM processing the received transmission symbols to provide received symbols, and (2) processing the received symbols based on the selected transmission mode(s) to obtain recovered symbols. The recovered symbols are estimates of the modulation symbols transmitted by the access point. The processing by receive processor 260 is described in further detail below.

[1041] A receive (RX) data processor 262 then symbol demaps, deinterleaves, and decodes the recovered symbols to obtain the user-specific data and signaling transmitted on the downlink for the terminal. The processing by receive processor 260 and RX data processor 262 is complementary to that performed by transmit processor 220 and TX data processor 210, respectively, at the access point.

[1042] On the uplink, at terminal 106, various types of traffic data such as user-specific data from a data source 276, signaling, and so on are provided to a TX data processor 278. Processor 278 codes the different types of traffic data in accordance with their respective coding schemes to provide coded data and further interleaves the coded data. A modulator 280 then symbol maps the interleaved data to provide modulated data, which is provided to one or more transmitters 254. OFDM may or may not be used for the uplink data transmission, depending on the system design. Each transmitter 254 conditions the received modulated data to generate a respective uplink modulated signal, which is then transmitted via an associated antenna 252 to the access point.

[1043] At access point 104, the uplink modulated signals from one or more terminals are received by antennas 224. The received signal from each antenna 224 is provided to a receiver 222, which conditions and digitizes the received signal to provide a respective stream of samples. The sample streams from all receivers 222 are then processed by a demodulator 240 and further decoded (if necessary) by an RX data processor 242 to recover the data transmitted by the terminals.

[1044] Controllers 230 and 270 direct the operation at the access point and the terminal, respectively. Memories 232 and 272 provide storage for program codes and data used by controllers 230 and 270, respectively. Scheduler 234 schedules the data transmission on the downlink (and possibly the uplink) for the terminals.

[1045] For clarity, various transmit diversity schemes are specifically described below for downlink transmission. These schemes may also be used for uplink transmission, and this is within the scope of the invention. Also for clarity, in the following description, subscript "*i*" is used as an index for the receive antennas, subscript "*j*" is used as an index for the transmit antennas, and subscript "*k*" is used as an index for the subbands in the MIMO OFDM system.

Transmitter Unit

[1046] **FIG. 3** is a block diagram of a transmitter unit 300, which is an embodiment of the transmitter portion of access point 104. Transmitter unit 300 includes (1) a TX data processor 210a that receives and processes traffic and pilot data to provide modulation symbols and (2) a transmit processor 220a that further processes the modulation symbols to provide N_T streams of transmission symbols for the N_T transmit antennas. TX data processor 210a and transmit processor 220a are one embodiment of TX data processor 210 and transmit processor 220, respectively, in FIG. 2.

[1047] In the specific embodiment shown in FIG. 3, TX data processor 210a includes an encoder 312, a channel interleaver 314, and a symbol mapping element 316. Encoder 312 receives and codes the traffic data (i.e., the information bits) based on one or more coding schemes to provide coded bits. The coding increases the reliability of the data transmission.

[1048] In an embodiment, the user-specific data for each terminal and the data for each overhead channel may be considered as distinct data streams. The overhead

channels may include broadcast, paging, and other common channels intended to be received by all terminals. Multiple data streams may also be sent to a given terminal. Each data stream may be coded independently based on a specific coding scheme selected for that data stream. Thus, a number of independently coded data streams may be provided by encoder 312 for different overhead channels and terminals.

[1049] The specific coding scheme to be used for each data stream is determined by a coding control from controller 230. The coding scheme for each terminal may be selected, for example, based on feedback information received from the terminal. Each coding scheme may include any combination of forward error detection (FED) codes (e.g., a cyclic redundancy check (CRC) code) and forward error correction (FEC) codes (e.g., a convolutional code, a Turbo code, a block code, and so on). A coding scheme may also designate no coding at all. Binary or trellis-based codes may also be used for each data stream. Moreover, with convolutional and Turbo codes, puncturing may be used to adjust the code rate. More specifically, puncturing may be used to increase the code rate above the base code rate.

[1050] In a specific embodiment, the data for each data stream is initially partitioned into frames (or packets). For each frame, the data may be used to generate a set of CRC bits for the frame, which is then appended to the data. The data and CRC bits for each frame are then coded with either a convolutional code or a Turbo code to generate the coded data for the frame.

[1051] Channel interleaver 314 receives and interleaves the coded bits based on one or more interleaving schemes. Typically, each coding scheme is associated with a corresponding interleaving scheme. In this case, each independently coded data stream would be interleaved separately. The interleaving provides time diversity for the coded bits, permits each data stream to be transmitted based on an average SNR of the subbands and spatial subchannels used for the data stream, combats fading, and further removes correlation between coded bits used to form each modulation symbol.

[1052] With OFDM, the channel interleaver may be designed to distribute the coded data for each data stream over multiple subbands of a single OFDM symbol or possibly over multiple OFDM symbols. The objective of the channel interleaver is to randomize the coded data so that the likelihood of consecutive coded bits being corrupted by the communication channel is reduced. When the interleaving interval for a given data

stream spans a single OFDM symbol, the coded bits for the data stream are randomly distributed across the subbands used for the data stream to exploit frequency diversity. When the interleaving interval spans multiple OFDM symbols, the coded bits are randomly distributed across the data-carrying subbands and the multi-symbol interleaving interval to exploit both frequency and time diversity. For a wireless local area network (WLAN), the time diversity realized by interleaving over multiple OFDM symbols may not be significant if the minimum expected coherence time of the communication channel is many times longer than the interleaving interval.

[1053] Symbol mapping element 316 receives and maps the interleaved data in accordance with one or more modulation schemes to provide modulation symbols. A particular modulation scheme may be used for each data stream. The symbol mapping for each data stream may be achieved by grouping sets of q_m coded and interleaved bits to form data symbols (each of which may be a non-binary value), and mapping each data symbol to a point in a signal constellation corresponding to the modulation scheme selected for use for that data stream. The selected modulation scheme may be QPSK, M-PSK, M-QAM, or some other modulation scheme. Each mapped signal point is a complex value and corresponds to an M_m-ary modulation symbol, where M_m corresponds to the specific modulation scheme selected for data stream m and $M_m = 2^{q_m}$. Symbol mapping element 316 provides a stream of modulation symbols for each data stream. The modulation symbol streams for all data streams are collectively shown as modulation symbol stream $s(n)$ in FIG. 3.

[1054] Table 1 lists various coding and modulation schemes that may be used to achieve a range of spectral efficiencies (or bit rates) using convolutional and Turbo codes. Each bit rate (in unit of bits/sec/Hertz or bps/Hz) may be achieved using a specific combination of code rate and modulation scheme. For example, a bit rate of one-half may be achieved using a code rate of 1/2 and BPSK modulation, a bit rate of one may be achieved using a code rate of 1/2 and QPSK modulation, and so on.

[1055] In Table 1, BPSK, QPSK, 16-QAM, and 64-QAM are used for the listed bit rates. Other modulation schemes such as DPSK, 8-PSK, 32-QAM, 128-QAM, and so on, may also be used and are within the scope of the invention. DPSK (differential phase-shift keying) may be used when the communication channel is difficult to track since a coherence reference is not needed at the receiver to demodulate a DPSK

modulated signal. For OFDM, modulation may be performed on a per subband basis, and the modulation scheme to be used for each subband may be independently selected.

Table 1

Convolutional Code			**Turbo Code**		
Efficiency (bps/Hz)	Code rate	Modulation	Efficiency (bps/Hz)	Code rate	Modulation
0.5	1/2	BPSK	0.5	1/2	BPSK
1.0	1/2	QPSK	1.0	1/2	QPSK
1.5	3/4	QPSK	1.5	3/4	QPSK
2.0	1/2	16-QAM	2.0	1/2	16-QAM
2.67	2/3	16-QAM	2.5	5/8	16-QAM
3.0	3/4	16-QAM	3.0	3/4	16-QAM
3.5	7/8	16-QAM	3.5	7/12	64-QAM
4.0	2/3	64-QAM	4.0	2/3	64-QAM
4.5	3/4	64-QAM	4.5	3/4	64-QAM
5.0	5/6	64-QAM	5.0	5/6	64-QAM

Other combinations of code rates and modulation schemes may also be used to achieve the various bit rates, and this is also within the scope of the invention.

[1056] In the specific embodiment shown in FIG. 3, transmit processor 220a includes a TX diversity processor 320 and N_T OFDM modulators. Each OFDM modulator includes an inverse fast Fourier transform (IFFT) unit 330 and a cyclic prefix generator 332. TX diversity processor 320 receives and processes the modulation symbols from TX data processor 210a in accordance with one or more selected transmission modes to provide transmit symbols.

[1057] In an embodiment, TX diversity processor 320 further receives and multiplexes pilot symbols (i.e., pilot data) with the transmit symbols using frequency division multiplexing (FDM) in a subset of the available subbands. An example implementation of an FDM pilot transmission scheme is shown in Table 2. In this implementation, 64 subbands are available for the MIMO OFDM system, and subband indices ±7 and ±21 are used for pilot transmission. In alternative embodiments, the pilot symbols may be multiplexed with the transmit symbols using, for example, time

division multiplexing (TDM), code division multiplexing (CDM), or any combination of FDM, TDM, and CDM.

[1058] TX diversity processor 320 provides one transmit symbol stream to each OFDM modulator. The processing by TX diversity processor 320 is described in further detail below.

[1059] Each OFDM modulator receives a respective transmit symbol stream $x_j(n)$. Within each OFDM modulator, IFFT unit 330 groups each set of N_F transmit symbols in stream $x_j(n)$ to form a corresponding symbol vector, and converts the symbol vector into its time-domain representation (which is referred to as an OFDM symbol) using the inverse fast Fourier transform.

[1060] For each OFDM symbol, cyclic prefix generator 332 repeats a portion of the OFDM symbol to form a corresponding transmission symbol. The cyclic prefix ensures that the transmission symbol retains its orthogonal property in the presence of multipath delay spread, thereby improving performance against deleterious path effects such as channel dispersion caused by frequency selective fading. A fixed or an adjustable cyclic prefix may be used for each OFDM symbol. As a specific example of an adjustable cyclic prefix, a system may have a bandwidth of 20 MHz, a chip period of 50 nsec, and 64 subbands. For this system, each OFDM symbol would have a duration of 3.2 μsec (or 64×50 nsec). The cyclic prefix for each OFDM symbol may have a minimum length of 4 chips (200 nsec) and a maximum length of 16 chips (800 nsec), with an increment of 4 chips (200 nsec). Each transmission symbol would then have a duration ranging from 3.4 μsec to 4.0 μsec for cyclic prefixes of 200 nsec to 800 nsec, respectively.

[1061] Cyclic prefix generator 332 in each OFDM modulator provides a stream of transmission symbols to an associated transmitter 222. Each transmitter 222 receives and processes a respective transmission symbol stream to generate a downlink modulated signal, which is then transmitted from the associated antenna 224.

[1062] The coding and modulation for a MIMO OFDM system are described in further detail in the following U.S. patent applications:

- U.S. Patent Application Serial No. 09/993,087, entitled "Multiple-Access Multiple-Input Multiple-Output (MIMO) Communication System," filed November 6, 2001;
- U.S. Patent Application Serial No. 09/854,235, entitled "Method and Apparatus for Processing Data in a Multiple-Input Multiple-Output (MIMO) Communication System Utilizing Channel State Information," filed May 11, 2001;
- U.S. Patent Application Serial Nos. 09/826,481 and 09/956,449, both entitled "Method and Apparatus for Utilizing Channel State Information in a Wireless Communication System," respectively filed March 23, 2001 and September 18, 2001;
- U.S. Patent Application Serial No. 09/776,075, entitled "Coding Scheme for a Wireless Communication System," filed February 1, 2001; and
- U.S. Patent Application Serial No. 09/532,492, entitled "High Efficiency, High Performance Communications System Employing Multi-Carrier Modulation," filed March 30, 2000.

These patent applications are all assigned to the assignee of the present application and incorporated herein by reference.

[1063] The MIMO OFDM system may be designed to support a number of modes of operation for data transmission. These transmission modes include diversity transmission modes, spatial multiplexing transmission modes, and beam steering transmission modes.

[1064] The spatial multiplexing and beam steering modes may be used to achieve higher bit rates under certain favorable channel conditions. These transmission modes are described in further detail in U.S. Patent Application Serial No. 10/085,456, entitled "Multiple-Input, Multiple-Output (MIMO) Systems with Multiple Transmission Modes," filed February 26, 2002, assigned to the assignee of the present application and incorporated herein by reference.

[1065] The diversity transmission modes may be used to achieve higher reliability for certain data transmissions. For example, the diversity transmission modes may be used for overhead channels on the downlink, such as broadcast, paging, and other common channels. The diversity transmission modes may also be used for data

transmission (1) whenever the transmitter does not have adequate channel state information (CSI) for the communication channel, (2) when the channel conditions are sufficiently poor (e.g., under certain mobility conditions) and cannot support more spectrally efficient transmission modes, and (3) for other situations. When the diversity transmission modes are used for downlink data transmission to the terminals, the rate and/or power for each terminal may be controlled to improve performance. A number of diversity transmission modes may be supported and are described in further detail below.

[1066] The diversity transmission modes attempt to achieve transmit diversity by establishing orthogonality among the multiple signals transmitted from multiple transmit antennas. Orthogonality among the transmitted signals may be attained in frequency, time, space, or any combination thereof. Transmit diversity may be established via any one or combination of the following processing techniques:

- Frequency (or subband) diversity. The inherent orthogonality among the subbands provided by OFDM is used to provide diversity against frequency selective fading.
- Transmit diversity using orthogonal functions. Walsh functions or some other orthogonal functions are applied to OFDM symbols transmitted from multiple transmit antennas to establish orthogonality among the transmitted signals. This scheme is also referred to herein as the "Walsh diversity" scheme.
- Space time transmit diversity (STTD). Spatial orthogonality is established between pairs of transmit antennas while preserving the potential for high spectral efficiency offered by MIMO techniques.

[1067] In general, the frequency diversity scheme may be used to combat frequency selective fading and operates in the frequency and spatial dimensions. The Walsh diversity scheme and STTD scheme operate in the time and spatial dimensions.

[1068] For clarity, the processing techniques enumerated above and certain combinations thereof will be described for an example MIMO OFDM system. In this system, each access point is equipped with four antennas to transmit and receive data, and each terminal may be equipped with one or multiple antennas.

Frequency Diversity

[1069] FIG. 4 is a block diagram of an embodiment of a TX diversity processor 320a that may be used to implement the frequency diversity scheme. For OFDM, the subbands are inherently orthogonal to one another. Frequency diversity may be established by transmitting identical modulation symbols on multiple subbands.

[1070] As shown in FIG. 4, the modulation symbols, $s(n)$, from TX data processor 210 are provided to a symbol repetition unit 410. Unit 410 repeats each modulation symbol based on the (e.g., dual or quad) diversity to be provided for the modulation symbol. A demultiplexer 412 then receives the repeated symbols and pilot symbols and demultiplexes these symbols into N_T transmit symbol streams. The modulation symbols for each data stream may be transmitted on a respective group of one or more subbands assigned to that data stream. Some of the available subbands may be reserved for pilot transmission (e.g., using FDM). Alternatively, the pilot symbols may be transmitted along with the modulation symbols using TDM or CDM.

[1071] In general, it is desirable to transmit repeated symbols in subbands that are separated from each other by at least the coherence bandwidth of the communication channel. Moreover, the modulation symbols may be repeated over any number of subbands. A higher repetition factor corresponds to greater redundancy and improved likelihood of correct reception at the receiver at the expense of reduced efficiency.

[1072] For clarity, a specific implementation of the frequency diversity scheme is described below for a specific MIMO OFDM system that has some of the characteristics defined by the IEEE Standard 802.11a. The specifications for this IEEE standard are described in a document entitled "Part 11: Wireless LAN Medium Access Control (MAC) and Physical Layer (PHY) specifications: High-speed Physical Layer in the 5 GHz Band," September 1999, which is publicly available and incorporated herein by reference. This system has an OFDM waveform structure with 64 subbands. Of these 64 subbands, 48 subbands (with indices of $\pm\{1, \ldots, 6, 8, \ldots, 20, 22, \ldots, 26\}$) are used for data, 4 subbands (with indices of $\pm\{7, 21\}$) are used for pilot, the DC subband (with index of 0) is not used, and the remaining subbands are also not used and serve as guard subbands.

[1073] Table 2 shows a specific implementation for dual and quad frequency diversity for the system described above. For dual frequency diversity, each modulation

symbol is transmitted over two subbands that are separated by either 26 or 27 subbands. For quad frequency diversity, each modulation symbol is transmitted over four subbands that are separated by 13 or 14 subbands. Other frequency diversity schemes may also be implemented and are within the scope of the invention.

Table 2

Subband Indices	Dual Diversity	Quad Diversity	Subband Indices	Dual Diversity	Quad Diversity
-26	1	1	1	1	1
-25	2	2	2	2	2
-24	3	3	3	3	3
-23	4	4	4	4	4
-22	5	5	5	5	5
-21	pilot	pilot	6	6	6
-20	6	6	7	pilot	pilot
-19	7	7	8	7	7
-18	8	8	9	8	8
-17	9	9	10	9	9
-16	10	10	11	10	10
-15	11	11	12	11	11
-14	12	12	13	12	12
-13	13	1	14	13	1
-12	14	2	15	14	2
-11	15	3	16	15	3
-10	16	4	17	16	4
-9	17	5	18	17	5
-8	18	6	19	18	6
-7	pilot	pilot	20	19	7
-6	19	7	21	pilot	pilot
-5	20	8	22	21	8
-4	21	9	23	22	9
-3	22	10	24	23	10
-2	23	11	25	24	11
-1	24	12	26	25	12
0	DC	DC	-	-	-

[1074] The frequency diversity scheme may be used by a transmitter (e.g., a terminal) not equipped with multiple transmit antennas. In this case, one transmit symbol stream is provided by TX diversity processor 310a. Each modulation symbol in $s(n)$ may be repeated and transmitted on multiple subbands. For single-antenna terminals, frequency diversity may be used to provide robust performance in the presence of frequency selective fading.

[1075] The frequency diversity scheme may also be used when multiple transmit antennas are available. This may be achieved by transmitting the same modulation symbol from all transmit antennas on distinct subbands or subband groups. For example, in a four transmit antenna device, every fourth subband may be assigned to one of the transmit antennas. Each transmit antenna would then be associated with a different group of $N_F/4$ subbands. For quad frequency diversity, each modulation symbol would then be transmitted on a set of four subbands, one in each of the four subband groups, with each group being associated with a specific transmit antenna. The four subbands in the set may also be selected such that they are spaced as far apart as possible. For dual frequency diversity, each modulation may be transmitted on a set of two subbands, one in each of two subband groups. Other implementations for frequency diversity with multiple transmit antennas may also be contemplated, and this is within the scope of the invention. The frequency diversity scheme may also be used in combination with one or more other transmit diversity schemes, as described below.

Walsh Transmit Diversity

[1076] **FIG. 5** is a block diagram of an embodiment of a TX diversity processor 320b that may be used to implement the Walsh diversity scheme. For this diversity scheme, orthogonal functions (or codes) are used to establish time orthogonality, which may in turn be used to establish full transmit diversity across all transmit antennas. This is achieved by repeating the same modulation symbols across the transmit antennas, and time spreading these symbols with a different orthogonal function for each transmit antenna, as described below. In general, various orthogonal functions may be used such as Walsh functions, orthogonal variable spreading factor (OVSF) codes, and so on. For clarity, Walsh functions are used in the following description.

[1077] In the embodiment shown in FIG. 5, the modulation symbols, $s(n)$, from TX data processor 210 are provided to a demultiplexer 510, which demultiplexes the symbols into N_B modulation symbol substreams, one substream for each subband used for data transmission (i.e., each data-carrying subband). Each modulation symbol substream $s_k(n)$ is provided to a respective TX subband processor 520.

[1078] Within each TX subband processor 520, the modulation symbols in substream $s_k(n)$ are provided to N_T multipliers 524a through 524d for the N_T transmit antennas (where $N_T = 4$ for this example system). In the embodiment shown in FIG. 5, one modulation symbol s_k is provided to all four multipliers 524 for each 4-symbol period, which corresponds to a symbol rate of $(4\mathrm{T}_{\mathrm{OFDM}})^{-1}$. Each multiplier also receives a different Walsh function having four chips (i.e., $W_j^4 = \{w_{1j},\ w_{2j},\ w_{3j},\ w_{4j}\}$) and assigned to transmit antenna j associated with that multiplier. Each multiplier then multiplies the symbol s_k with the Walsh function W_j^4 and provides a sequence of four transmit symbols, $\{(s_k \cdot w_{1j}),\ (s_k \cdot w_{2j}),\ (s_k \cdot w_{3j}),\ \text{and}\ (s_k \cdot w_{4j})\}$, which is to be transmitted in four consecutive OFDM symbol periods on subband k of transmit antenna j. These four transmit symbols have the same magnitude as the original modulation symbol s_k. However, the sign of each transmit symbol in the sequence is determined by the sign of the Walsh chip used to generate that transmit symbol. The Walsh function is thus used to time-spread each modulation symbol over four symbol periods. The four multipliers 524a through 524d of each TX subband processor 520 provide four transmit symbol substreams to four buffers/multiplexers 530a through 530d, respectively.

[1079] Each buffer/multiplexer 530 receives pilot symbols and N_B transmit symbol substreams for N_B subbands from N_B TX subband processors 520a through 520f. Each unit 530 then multiplexes the transmit symbols and pilot symbols for each symbol period, and provides a stream of transmit symbols $x_j(n)$ to a corresponding IFFT unit 330. Each IFFT unit 330 receives and processes a respective transmit symbol stream $x_j(n)$ in the manner described above.

[1080] In the embodiment shown in FIG. 5, one modulation symbol is transmitted from all four transmit antennas on each of the N_B data-carrying subbands for each 4-

symbol period. When four transmit antennas are used for data transmission, the spectral efficiency achieved with the Walsh diversity scheme is identical to that achieved with the quad frequency diversity scheme whereby one modulation symbol is transmitted over four data-carrying subbands for each symbol period. In the Walsh diversity scheme with four transmit antennas, the duration or length of the Walsh functions is four OFDM symbols (as designated by the superscript in W_j^4). Since the information in each modulation symbol is distributed over four successive OFDM symbols, the demodulation at the receiver is performed based on four consecutive received OFDM symbols.

[1081] In an alternative embodiment, increased spectral efficiency may be achieved by transmitting distinct modulation symbols (instead of the same modulation symbol) on each transmit antenna. For example, demultiplexer 510 may be designed to provide four distinct modulation symbols, s_1, s_2, s_3, and s_4, to multipliers 524a through 524d for each 4-symbol period. Each multiplier 524 would then multiply a different modulation symbol with its Walsh function to provide a different sequence of four transmit symbols. The spectral efficiency for this embodiment would then be four times that of the embodiment shown in FIG. 5. As another example, demultiplexer 510 may be designed to provide two distinct modulation symbols (e.g., s_1 to multipliers 524a and 524b and s_2 to multipliers 524c and 524d) for each 4-symbol period.

Space-Time Transmit Diversity (STTD)

[1082] Space-time transmit diversity (STTD) supports simultaneous transmission of effectively two independent symbol streams on two transmit antennas while maintaining orthogonality at the receiver. An STTD scheme may thus provide higher spectral efficiency over the Walsh transmit diversity scheme shown in FIG. 5.

[1083] The STTD scheme operates as follows. Suppose that two modulation symbols, denoted as s_1 and s_2, are to be transmitted on a given subband. The transmitter generates two vectors, $\underline{\mathbf{x}}_1 = [s_1 \;\; s_2^*]^T$ and $\underline{\mathbf{x}}_2 = [s_2 \;\; -s_1^*]^T$. Each vector includes two elements that are to be transmitted sequentially in two symbol periods from a respective transmit antenna (i.e., vector $\underline{\mathbf{x}}_1$ is transmitted from antenna 1 and vector $\underline{\mathbf{x}}_2$ is transmitted from antenna 2).

[1084] If the receiver includes a single receive antenna, then the received signal may be expressed in matrix form as:

$$\begin{bmatrix} r_1 \\ r_2 \end{bmatrix} = \begin{bmatrix} h_1 s_1 + h_2 s_2 \\ h_1 s_2^* - h_2 s_1^* \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \end{bmatrix} , \qquad \text{Eq (1)}$$

where r_1 and r_2 are two symbols received in two consecutive symbol periods at the receiver;

h_1 and h_2 are the path gains from the two transmit antennas to the receive antenna for the subband under consideration, where the path gains are assumed to be constant over the subband and static over the 2-symbol period; and

n_1 and n_2 are the noise associated with the two received symbols r_1 and r_2.

[1085] The receiver may then derive estimates of the two transmitted symbols, s_1 and s_2, as follows:

$$\hat{s}_1 = h_1^* r_1 - h_2 r_2^* = (|h_1|^2 + |h_2|^2) s_1 + h_1^* n_1 - h_2 n_2 \text{ , and} \qquad \text{Eq (2)}$$

$$\hat{s}_2 = h_2^* r_1 + h_1 r_2^* = (|h_1|^2 + |h_2|^2) s_2 + h_2^* n_1 + h_1 n_2 \text{ .}$$

[1086] In an alternative implementation, the transmitter may generate two vectors, $\underline{\mathbf{x}}_1 = [s_1 \;\; s_2]^T$ and $\underline{\mathbf{x}}_2 = [-s_2^* \;\; s_1^*]^T$, with the elements of these two vectors being transmitted sequentially in two symbol periods from the two transmit antennas. The received signal may then be expressed as:

$$\begin{bmatrix} r_1 \\ r_2 \end{bmatrix} = \begin{bmatrix} h_1 s_1 - h_2 s_2^* \\ h_1 s_2 + h_2 s_1^* \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \end{bmatrix} .$$

The receiver may then derive estimates of the two transmitted symbols as follows:

$$\hat{s}_1 = h_1^* r_1 + h_2 r_2^* = (|h_1|^2 + |h_2|^2) s_1 + h_1^* n_1 + h_2 n_2 \text{ , and}$$

$$\hat{s}_2 = -h_2 r_1^* + h_1^* r_2 = (|h_1|^2 + |h_2|^2) s_2 - h_2 n_1 + h_1^* n_2 \text{ .}$$

[1087] When two transmit antennas are employed for data transmission, the STTD scheme is twice as spectrally efficient as both the dual frequency diversity scheme and the Walsh diversity scheme with two transmit antennas. The STTD scheme effectively transmits one independent modulation symbol per subband over the two transmit antennas in each symbol period, whereas the dual frequency diversity scheme transmits only a single modulation symbol per two subbands in each symbol period and the Walsh diversity scheme transmits only a single modulation symbol on each subband in two symbol periods. Since the information in each modulation symbol is distributed over two successive OFDM symbols for the STTD scheme, the demodulation at the receiver is performed based on two consecutive received OFDM symbols.

[1088] **FIG. 6** is a block diagram of an embodiment of a TX diversity processor 320c that may be used to implement the STTD scheme. In this embodiment, the modulation symbols, $s(n)$, from TX data processor 210 are provided to a demultiplexer 610, which demultiplexes the symbols into $2N_B$ modulation symbol substreams, two substreams for each data-carrying subband. Each pair of modulation symbol substreams is provided to a respective TX subband processor 620. Each modulation symbol substream includes one modulation symbol for each 2-symbol period, which corresponds to a symbol rate of $(2T_{OFDM})^{-1}$.

[1089] Within each TX subband processor 620, the pair of modulation symbol substreams is provided to a space-time encoder 622. For each pair of modulation symbols in the two substreams, space-time encoder 622 provides two vectors, $\underline{\mathbf{x}}_1 = [s_1 \;\; s_2^*]^T$ and $\underline{\mathbf{x}}_2 = [s_2 \;\; -s_1^*]^T$, with each vector including two transmit symbols to be transmitted in two symbol periods. The two transmit symbols in each vector have the same magnitude as the original modulation symbols, s_1 and s_2. However, each transmit symbol may be rotated in phase relative to the original modulation symbol. Each TX subband processor 620 thus provides two transmit symbol substreams to two buffers/multiplexers 630a and 630b, respectively.

[1090] Each buffer/multiplexer 630 receives pilot symbols and N_B transmit symbol substreams from N_B TX subband processors 620a through 620f, multiplexes the transmit symbols and pilot symbols for each symbol period, and provides a stream of transmit

symbols $x_j(n)$ to a corresponding IFFT unit 330. Each IFFT unit 330 then processes a respective transmit symbol stream in the manner described above.

[1091] The STTD scheme is described in further detail by S.M. Alamouti in a paper entitled "A Simple Transmit Diversity Technique for Wireless Communications," IEEE Journal on Selected Areas in Communications, Vol. 16, No. 8, October 1998, pgs. 1451-1458, which is incorporated herein by reference. The STTD scheme is also described in further detail in U.S. Patent Application Serial No. 09/737,602, entitled "Method and System for Increased Bandwidth Efficiency in Multiple Input - Multiple Output Channels," filed January 5, 2001, assigned to the assignee of the present application and incorporated herein by reference.

Walsh-STTD

[1092] A Walsh-STTD scheme employs a combination of Walsh diversity and STTD described above. The Walsh-STTD scheme may be used in systems with more than two transmit antennas. For a Walsh-STTD with repeated symbols scheme (which is also referred to as the repeated Walsh-STTD scheme), two transmit vectors $\underline{\mathbf{x}}_1$ and $\underline{\mathbf{x}}_2$ are generated for each pair of modulation symbols to be transmitted on a given subband from two transmit antennas, as described above for FIG. 6. These two transmit vectors are also repeated across multiple pairs of transmit antennas using Walsh functions to achieve orthogonality across the transmit antenna pairs and to provide additional transmit diversity.

[1093] **FIG. 7** is a block diagram of an embodiment of a TX diversity processor 320d that may be used to implement the repeated Walsh-STTD scheme. The modulation symbols, $s(n)$, from TX data processor 210 are provided to a demultiplexer 710, which demultiplexes the symbols into $2N_B$ modulation symbol substreams, two substreams for each data-carrying subband. Each modulation symbol substream includes one modulation symbol for each 4-symbol period, which corresponds to a symbol rate of $(4T_{OFDM})^{-1}$. Each pair of modulation symbol substreams is provided to a respective TX subband processor 720.

[1094] A space-time encoder 722 within each TX subband processor 720 receives the pair of modulation symbol substreams and, for each 4-symbol period, forms a pair

of modulation symbols $\{s_1$ and $s_2\}$, with one symbol coming from each of the two substreams. The pair of modulation symbols $\{s_1$ and $s_2\}$ is then used to form two vectors, $\underline{\mathbf{x}}_1 = [s_1 \;\; s_2^*]^T$ and $\underline{\mathbf{x}}_2 = [s_2 \;\; -s_1^*]^T$, with each vector spanning a 4-symbol period. Space-time encoder 722 provides the first vector $\underline{\mathbf{x}}_1$ to multipliers 724a and 724c and the second vector $\underline{\mathbf{x}}_2$ to multipliers 724b and 724d. Multipliers 724a and 724b each also receive a Walsh function having two chips (i.e., $W_1^2 = \{w_{11}, \; w_{21}\}$) and assigned to transmit antennas 1 and 2. Similarly, multipliers 724c and 724d each also receive a Walsh function W_2^2 having two chips and assigned to transmit antennas 3 and 4. Each multiplier 724 then multiplies each symbol in its vector $\underline{\mathbf{x}}_j$ with its Walsh function to provide two transmit symbols to be transmitted in two consecutive symbol periods on subband k of transmit antenna j.

[1095] In particular, multiplier 724a multiplies each symbol in vector $\underline{\mathbf{x}}_1$ with the Walsh function W_1^2 and provides a sequence of four transmit symbols, $\{(s_1 \cdot w_{11}),\ (s_1 \cdot w_{21}),\ (s_2^* \cdot w_{11}),$ and $(s_2^* \cdot w_{21})\}$, which is to be transmitted in four consecutive symbol periods. Multiplier 724b multiplies each symbol in vector $\underline{\mathbf{x}}_2$ with the Walsh function W_1^2 and provides a sequence of four transmit symbols, $\{(s_2 \cdot w_{11}),\ (s_2 \cdot w_{21}),\ (-s_1^* \cdot w_{11}),$ and $(-s_1^* \cdot w_{21})\}$. Multiplier 724c multiplies each symbol in vector $\underline{\mathbf{x}}_1$ with the Walsh function W_2^2 and provides a sequence of four transmit symbols, $\{(s_1 \cdot w_{12}),\ (s_1 \cdot w_{22}),\ (s_2^* \cdot w_{12}),$ and $(s_2^* \cdot w_{22})\}$. And multiplier 724d multiplies each symbol in vector $\underline{\mathbf{x}}_2$ with the Walsh function W_2^2 and provides a sequence of four transmit symbols, $\{(s_2 \cdot w_{12}),\ (s_2 \cdot w_{22}),\ (-s_1^* \cdot w_{12}),$ and $(-s_1^* \cdot w_{22})\}$. The Walsh function is thus used to time-spread each symbol or element in the vector $\underline{\mathbf{x}}$ over two symbol periods. The four multipliers 724a through 724d of each TX subband processor 720 provide four transmit symbol substreams to four buffers/multiplexers 730a through 730d, respectively.

[1096] Each buffer/multiplexer 730 receives pilot symbols and N_B transmit symbol substreams from N_B TX subband processors 720a through 720f, multiplexes the pilot and transmit symbols for each symbol period, and provides a stream of transmit

symbols $x_j(n)$ to a corresponding IFFT unit 330. The subsequent processing is as described above.

[1097] The repeated Walsh-STTD scheme shown in FIG. 7 (with four transmit antennas) has the same spectral efficiency as the STTD scheme shown in FIG. 6 and twice the spectral efficiency of the Walsh diversity scheme shown in FIG. 5. However, additional diversity is provided by this Walsh-STTD scheme by transmitting repeated symbols over multiple pairs of transmit antennas. The Walsh-STTD processing provides full transmit diversity (per subband) for the signals transmitted from all transmit antennas.

[1098] **FIG. 8** is a block diagram of an embodiment of a TX diversity processor 320e that may be used to implement a Walsh-STTD without repeated symbols scheme (which is also referred to as the non-repeated Walsh-STTD scheme). This scheme may be used to increase spectral efficiency at the expense of less diversity than the scheme shown in FIG. 7. As shown in FIG. 8, the modulation symbols $s(n)$ are provided to a demultiplexer 810, which demultiplexes the symbols into $4N_B$ modulation symbol substreams, four substreams for each data-carrying subband. Each set of four modulation symbol substreams is provided to a respective TX subband processor 820.

[1099] Within each TX subband processor 820, a space-time encoder 822a receives the first pair of modulation symbol substreams and a space-time encoder 822b receives the second pair of modulation symbol substreams. For each pair of modulation symbols in the two substreams in the first pair, space-time encoder 822a provides two vectors $\underline{\mathbf{x}}_1 = [s_1 \;\; s_2^*]^T$ and $\underline{\mathbf{x}}_2 = [s_2 \;\; -s_1^*]^T$ to multipliers 824a and 824b, respectively. Similarly, for each pair of modulation symbols in the two substreams in the second pair, space-time encoder 822b provides two vectors $\underline{\mathbf{x}}_3 = [s_3 \;\; s_4^*]^T$ and $\underline{\mathbf{x}}_4 = [s_4 \;\; -s_3^*]^T$ to multipliers 824c and 824d, respectively

[1100] Multipliers 824a and 824b each also receive Walsh function W_1^2, and multipliers 824c and 824d each also receive Walsh function W_2^2. Each multiplier 824 then multiplies each symbol in its vector $\underline{\mathbf{x}}_j$ with its Walsh function to provide two transmit symbols to be transmitted in two consecutive symbol periods on subband k of transmit antenna j. The four multipliers 824a through 824d of each TX subband

processor 820 provide four transmit symbol substreams to four buffers/multiplexers 830a through 830d, respectively.

[1101] Each buffer/multiplexer 830 receives pilot symbols and N_B transmit symbol substreams from N_B TX subband processors 820a through 820f, multiplexes the pilot symbols and transmit symbols for each symbol period, and provides a stream of transmit symbols $x_j(n)$ to a corresponding IFFT unit 330. The subsequent processing is as described above.

[1102] The non-repeated Walsh-STTD scheme shown in FIG. 8 (with four transmit antennas) has twice the spectral efficiency as the repeated Walsh-STTD scheme shown in FIG. 7. The same processing may be extended to a system with any number of transmit antenna pairs. Instead of repeating the two transmit vectors across the pairs of transmit antennas, each transmit antenna pair may be used to transmit independent symbol streams. This results in greater spectral efficiency at the possible expense of diversity performance. Some of this diversity may be recovered by the use of forward error correction (FEC) code.

[1103] The Walsh-STTD scheme is also described in further detail in the aforementioned U.S. Patent Application Serial No. 09/737,602.

Frequency-STTD

[1104] A frequency-STTD scheme employs a combination of frequency diversity and STTD. The frequency-STTD scheme may also employ antenna diversity for systems with more than one pair of transmit antennas. For the frequency-STTD scheme, each modulation symbol is transmitted on multiple (e.g., two) subbands and provided to multiple TX subband processors. The subbands to be used for each modulation symbol may be selected such that they are spaced as far apart as possible (e.g., as shown in Table 1) or based on some other subband assignment scheme. If four transmit antennas are available, then for each subband two pairs of modulation symbols are processed using STTD. The first pair of modulation symbols is transmitted from the first pair of antennas (e.g., transmit antennas 1 and 2), and the second pair of modulation symbols is transmitted from the second pair of antennas (e.g., transmit antennas 3 and 4).

[1105] Each modulation symbol is thus transmitted on multiple subbands and over multiple transmit antennas. For clarity, the processing for a given modulation symbol s_a for a system with four transmit antennas and using dual frequency diversity may be performed as follows. Modulation symbol s_a is initially provided to two TX subband processors (e.g., for subbands k and $k + N_F/2$). In subband k, modulation symbol s_a is processed with another modulation symbol s_b using STTD to form two vectors, $\underline{\mathbf{x}}_1 = [s_a \;\; s_b^*]^T$ and $\underline{\mathbf{x}}_2 = [s_b \;\; -s_a^*]^T$, which are transmitted from transmit antennas 1 and 2, respectively. In subband $k + N_F/2$, modulation symbol s_a is processed with another modulation symbol s_c using STTD to form two vectors, $\underline{\mathbf{x}}_3 = [s_a \;\; s_c^*]^T$ and $\underline{\mathbf{x}}_4 = [s_c \;\; -s_a^*]^T$, which are transmitted from transmit antennas 3 and 4, respectively. Modulation symbol s_c may be the same as modulation symbol s_b or a different modulation symbol.

[1106] For the above implementation of the frequency-STTD scheme, the modulation symbol in each subband has two orders of transmit diversity provided by the STTD processing. Each modulation symbol to be transmitted has four orders of transmit diversity plus some frequency diversity provided by the use of two subbands and STTD. This frequency-STTD scheme has the same spectral efficiency as the repeated Walsh-STTD scheme. However, the total transmission time for each modulation symbol is two symbol periods with the frequency-STTD scheme, which is half the total transmission time for each modulation symbol with the Walsh-STTD scheme, since Walsh processing is not performed by the frequency-STTD scheme.

[1107] In one embodiment of the frequency-STTD scheme, all subbands are used by each pair of transmit antennas for data transmission. For quad diversity, each modulation symbol is provided to two subbands for two transmit antenna pairs, as described above. In another embodiment of the frequency-STTD scheme, each pair of transmit antennas is assigned a different subband group for data transmission. For example, in a device with two pairs of transmit antennas, every other subband may be assigned to one transmit antenna pair. Each transmit antenna pair would then be associated with a different group of $N_F/2$ subbands. For quad diversity, each modulation symbol would then be transmitted on two subbands, one in each of the two

subband groups, with each group being associated with a specific transmit antenna pair. The two subbands used for each modulation symbol may be selected such that they are spaced as far apart as possible. Other implementations for frequency-STTD diversity with multiple transmit antenna pairs may also be contemplated, and this is within the scope of the invention.

[1108] As illustrated by the above, various diversity schemes may be implemented using various processing techniques described herein. For clarity, specific implementations of various diversity schemes have been described above for a specific system. Variations of these diversity schemes may also be implemented, and this is within the scope of the invention.

[1109] Moreover, other diversity schemes may also be implemented based on other combinations of the processing techniques described herein, and this is also within the scope of the invention. For example, another diversity scheme may utilize frequency diversity and Walsh transmit diversity, and yet another diversity scheme may utilize frequency diversity, Walsh diversity, and STTD.

Diversity Transmission Modes

[1110] A number of diversity transmission modes may be implemented using the transmit processing schemes described above. These diversity transmission modes may include the following:

- Frequency diversity transmission mode - employs only frequency diversity (e.g., dual, quad, or some other integer multiple frequency diversity).
- Walsh diversity transmission mode - employs only Walsh transmit diversity.
- STTD transmission mode - employs only STTD.
- Walsh-STTD transmission mode - employs both Walsh transmit diversity and STTD, with repeated or non-repeated symbols.
- Frequency-STTD transmission mode - employs frequency diversity and STTD.
- Frequency-Walsh transmission mode - employs frequency diversity and Walsh transmit diversity.

- Frequency-Walsh-STTD transmission mode - employs frequency diversity. Walsh transmit diversity, and STTD.

[1111] The diversity transmission modes may be used for data transmission between the access points and terminals. The specific transmission mode to use for a given data stream may be dependent on various factors such as (1) the type of data being transmitted (e.g., whether common for all terminals or user-specific for a particular terminal), (2) the number of antennas available at the transmitter and receiver, (3) the channel conditions, (4) the requirements of the data transmission (e.g., the required packet error rate), and so on.

[1112] Each access point in the system may be equipped with, for example, four antennas for data transmission and reception. Each terminal may be equipped with one, two, four, or some other number of antennas for data transmission and reception. Default diversity transmission modes may be defined and used for each terminal type. In a specific embodiment, the following diversity transmission modes are used as default:

- Single-antenna terminals – use frequency diversity transmission mode with dual or quad diversity.
- Dual-antenna terminals – use STTD transmission mode for dual diversity and frequency-STTD transmission mode for quad diversity.
- Quad-antenna terminals – use STTD transmission mode for dual diversity and Walsh-STTD transmission mode for quad diversity.

Other diversity transmission modes may also be selected as the default modes, and this is within the scope of the invention.

[1113] The diversity transmission modes may also be used to increase the reliability of data transmission on overhead channels intended to be received by all terminals in the system. In an embodiment, a specific diversity transmission mode is used for the broadcast channel, and this mode is known *a priori* by all terminals in the system (i.e., no signaling is required to identify the transmission mode used for the broadcast channel). In this way, the terminals are able to process and recover the data transmitted on the broadcast channel. The transmission modes used for other overhead channels may be fixed or dynamically selected. In one dynamic selection scheme, the system

determines which transmission mode is the most reliable (and spectrally efficient) to use for each of the remaining overhead channels based on the mix of terminals being served. The transmission modes selected for use for these overhead channels and other configuration information may be signaled to the terminals, for example, via the broadcast channel.

[1114] With OFDM, the subbands may be treated as distinct transmission channels, and the same or different diversity transmission modes may be used for the subbands. For example, one diversity transmission mode may be used for all data-carrying subbands, or a separate diversity transmission mode may be selected for each data-carrying subband. Moreover, for a given subband, it may be possible to use different diversity transmission modes for different sets of transmit antennas.

[1115] In general, each data stream (whether for an overhead channel or a specific receiver device) may be coded and modulated based on the coding and modulation schemes selected for that data stream to provide modulation symbols. The modulation symbols are then further processed based on the diversity transmission mode selected for that data stream to provide transmit symbols. The transmit symbols are further processed and transmitted on a group of one or more subbands from a set of one or more transmit antennas designated to be used for that data stream.

Receiver Unit

[1116] **FIG. 9** is a block diagram of a receiver unit 900, which is an embodiment of the receiver portion of a multi-antenna terminal 106. The downlink modulated signals from access point 104 are received by antennas 252a through 252r, and the received signal from each antenna is provided to a respective receiver 254. Each receiver 254 processes (e.g., conditions, digitizes, and data demodulates) the received signal to provide a stream of received transmission symbols, which is then provided to a respective OFDM demodulator within a receive processor 260a.

[1117] Each OFDM demodulator includes a cyclic prefix removal unit 912 and a fast Fourier transform (FFT) unit 914. Unit 912 removes the cyclic prefix that had been appended in each transmission symbol to provide a corresponding received OFDM symbol. The cyclic prefix removal may be performed by determining a set of N_A samples corresponding to each received transmission symbol and selecting a subset of

these N_A samples as the set of N_F samples for the received OFDM symbol. FFT 914 then transforms each received OFDM symbol (or each set of N_F samples) using the fast Fourier transform to provide a vector of N_F received symbols for the N_F subbands. FFT units 914a through 914r provide N_R received symbol streams, $r_1(n)$ through $r_{N_R}(n)$, to an RX diversity processor 920.

[1118] RX diversity processor 920 performs diversity processing on the N_R received symbol streams to provide recovered symbols, $\hat{s}(n)$, which are estimates of the modulation symbols, $s(n)$, sent by the transmitter. The processing to be performed by RX diversity processor 920 is dependent on the transmission mode used for each data stream to be recovered, as indicated by the transmission mode control. RX diversity processor 920 is described in further detail below.

[1119] RX diversity processor 920 provides the recovered symbols, $\hat{s}(n)$, for all data streams to be recovered to an RX data processor 262a, which is an embodiment of RX data processor 262 in FIG. 2. Within processor 262a, a symbol demapping element 942 demodulates the recovered symbols for each data stream in accordance with a demodulation scheme that is complementary to the modulation scheme used for the data stream. A channel deinterleaver 944 then deinterleaves the demodulated data in a manner complementary to the interleaving performed at the transmitter for the data stream, and the deinterleaved data is further decoded by a decoder 946 in a manner complementary to the coding performed at the transmitter. For example, a Turbo decoder or a Viterbi decoder may be used for decoder 946 if Turbo or convolutional coding, respectively, is performed at the transmitter. The decoded data from decoder 946 represents an estimate of the transmitted data being recovered. Decoder 946 may also provide the status of each received packet (e.g., indicating whether it was received correctly or in error).

[1120] In the embodiment shown in FIG. 9, a channel estimator 950 estimates various channel characteristics such as the channel response and the noise variance (e.g., based on recovered pilot symbols) and provides these estimates to controller 270. Controller 270 may be designed to perform various functions related to diversity processing at the receiver. For example, controller 270 may determine the diversity transmission mode used for each data stream to be recovered and may further direct the operation of RX diversity processor 920.

[1121] **FIG. 10** is a block diagram of an embodiment of an RX diversity processor 920x, which may be used for a multi-antenna receiver device. In this embodiment, the N_R received symbol streams for the N_R receive antennas are provided to N_R RX antenna processors 1020a through 1020r. Each RX antenna processor 1020 processes a respective received symbol stream, $r_i(n)$, and provides a corresponding recovered symbol stream, $\hat{s}_i(n)$, for the associated receive antenna. In an alternative embodiment, one or more RX antenna processors 1020 are time shared and used to process all N_R received symbol streams.

[1122] A combiner 1030 then receives and combines the N_R recovered symbol streams from the N_R RX antenna processors 1020a through 1020r to provide a single recovered symbol stream, $\hat{s}(n)$. The combining may be performed on a symbol-by-symbol basis. In an embodiment, for a given subband k, the N_R recovered symbols from the N_R receive antennas for each symbol period (which are denoted as $\{\hat{s}_{ki}\}$, for $i = (1, 2, \ldots, N_R)$) are initially scaled by N_R weights assigned to the N_R receive antennas. The N_R scaled symbols are then summed to provide the recovered symbol, $\hat{s}_k$, for subband k. The weights may be selected to achieve maximal-ratio combining, and may be determined based on the signal quality (e.g., SNR) associated with the receive antennas. The scaling with the weights may also be performed via an automatic gain control (AGC) loop maintained for each receive antenna, as is known in the art.

[1123] For a single-antenna receiver device, there is only one received symbol stream. In this case, only one RX antenna processor 1020 is needed. A design for RX antenna processor 1020 is described in further detail below.

[1124] The recovered symbol steam, $\hat{s}(n)$, provided by combiner 1030 may include the recovered symbols for all data streams transmitted by the transmitter. Alternatively, the steam $\hat{s}(n)$ may include only the recovered symbols for one or more data streams to be recovered by the receiver device.

[1125] **FIG. 11** is a block diagram of an RX antenna processor 1020x that may be used to perform the receive processing for the Walsh diversity scheme shown in FIG. 5. RX antenna processor 1020x processes the received symbol stream $r_i(n)$ for one receive antenna and may be used for each of RX antenna processors 1020a through 1020r in FIG. 10.

[1126] In the embodiment shown in FIG. 11, the received symbol stream $r_i(n)$ is provided to a demultiplexer 1110, which demultiplexes the received symbols in $r_i(n)$ into N_B substreams of received symbols (which are denoted as r_1 through r_{N_B}, where the index i has been dropped for simplicity), one substream for each data-carrying subband. Each received symbol substream r_k is then provided to a respective RX subband processor 1120.

[1127] Each RX subband processor 1120 includes a number of receive processing paths, one path for each transmit antenna used for data transmission (four receive processing paths are shown in FIG. 11 for four transmit antennas). For each processing path, the received symbols in the substream are provided to a multiplier 1122 that also receives a scaled Walsh function $\hat{h}^*_{kj}(W^4_j)^*$, where $\hat{h}^*_{kj}$ is the complex-conjugated channel response estimate between transmit antenna j (which is associated with that multiplier) and the receive antenna for subband k, and $(W^4_j)^*$ is the complex-conjugated Walsh function assigned to transmit antenna j. Each multiplier 1122 then multiplies the received symbols with the scaled Walsh function and provides the results to an associated integrator 1124. Integrator 1124 then integrates the multiplier results over the length of the Walsh function (or four symbol periods) and provides the integrated output to a summer 1126. One received symbol is provided to multiplier 1122 for each symbol period (i.e., rate $= (T_{OFDM})^{-1}$) and integrator 1124 provides one integrated output for each 4-symbol period (i.e., rate $= (4T_{OFDM})^{-1}$).

[1128] For each 4-symbol period, summer 1126 combines the four outputs from integrators 1124a through 1124d to provide a recovered symbol, $\hat{s}_k$, for subband k, which is an estimate of the modulation symbol, s_k, transmitted in that subband. For each 4-symbol period, RX subband processors 1120a through 1120f provide N_B recovered symbols, $\hat{s}_1$ through $\hat{s}_{N_B}$, for the N_B data-carrying subbands.

[1129] A multiplexer 1140 receives the recovered symbols from RX subband processors 1120a through 1120f and multiplexes these symbols into a recovered symbol stream, $\hat{s}_i(n)$, for receive antenna i.

[1130] **FIG. 12** is a block diagram of an RX subband processor 1120x that may be used to perform the receive processing for the Walsh-STTD schemes shown in FIGS. 7

and 8. RX subband processor 1120x processes one received symbol substream r_k for one subband of one receive antenna and may be used for each of RX subband processors 1120a through 1120f in FIG. 11.

[1131] In the embodiment shown in FIG. 12, the received symbols in substream r_k are provided to two receive processing paths, one path for each transmit antenna pair used for data transmission (two receive processing paths are shown in FIG. 12 for four transmit antennas). For each processing path, the received symbols are provided to a multiplier 1222 that also receives a complex-conjugated Walsh function $(W_j^2)^*$ assigned to the transmit antenna pair being processed by that path. Each multiplier 1222 then multiplies the received symbols with the Walsh function and provides the results to an associated integrator 1224. Integrator 1224 then integrates the multiplier results over the length of the Walsh function (or two symbol periods) and provides the integrated output to a delay element 1226 and a unit 1228. One received symbol is provided to multiplier 1222 for each symbol period (i.e., $\text{rate} = (T_{OFDM})^{-1}$) and integrator 1224 provides one integrated output for each 2-symbol period (i.e., $\text{rate} = (2T_{OFDM})^{-1}$).

[1132] Referring back to FIG. 8, for the non-repeated Walsh-STTD scheme, four modulation symbols $\{s_{k1}$, s_{k2}, s_{k3}, and $s_{k4}\}$ are transmitted over two transmit antenna pairs in four symbol periods for subband *k* (where the index *k* is used to denote subband *k*). The symbol pair $\{s_{k1}$ and $s_{k2}\}$ is transmitted over the first transmit antenna pair, and the symbol pair $\{s_{k3}$ and $s_{k4}\}$ is transmitted over the second transmit antenna pair. Each modulation symbol is transmitted in two symbol periods using the 2-chip Walsh function assigned to the transmit antenna pair.

[1133] Referring back to FIG. 12, the complementary processing is performed at the receiver to recover the modulation symbols. For each 4-symbol period corresponding to a new symbol pair transmitted from each transmit antenna pair for subband *k*, integrator 1224 provides a received symbol pair $\{r_{k1}$ and $r_{k2}\}$. Delay element 1226 then provides a delay of two symbol periods (i.e., $T_W = 2T_{OFDM}$, which is the length of the Walsh function) for the first symbol (i.e., r_{k1}) in the pair, and unit 1228 provides the complex-conjugate of the second symbol (i.e., r_{k2}^*) in the pair.

[1134] Multipliers 1230a through 1230d and summers 1232a and 1232b then collectively perform the computations shown in equation (2) for the first transmit antenna pair. In particular, multiplier 1230a multiplies the symbol r_{k1} with the channel response estimate $\hat{h}_{k1}^*$, multiplier 1230b multiplies the symbol r_{k2}^* with the channel response estimate $\hat{h}_{k2}$, multiplier 1230c multiplies the symbol r_{k1} with the channel response estimate $\hat{h}_{k2}^*$, and multiplier 1230d multiplies the symbol r_{k2}^* with the channel response estimate $\hat{h}_{k1}$, where $\hat{h}_{kj}$ is an estimate of the channel response from transmit antenna j to the receive antenna for subband k. Summer 1232a then subtracts the output of multiplier 1230b from the output of multiplier 1230a to provide an estimate, $\hat{s}_{k1}$, of the first modulation symbol in the pair $\{s_{k1}$ and $s_{k2}\}$. Summer 1232b adds the output of multiplier 1230c with the output of multiplier 1230d to provide an estimate, $\hat{s}_{k2}$, of the second modulation symbol in the pair.

[1135] The processing by the second path for the second transmit antenna pair is similar to that described above for the first path. However, the channel response estimates, $\hat{h}_{k3}$ and $\hat{h}_{k4}$, for the second pair of transmit antennas for subband k are used for the second processing path. For each 4-symbol period, the second processing path provides the symbol estimates $\hat{s}_{k3}$ and $\hat{s}_{k4}$ for the pair of modulation symbols $\{s_{k3}$ and $s_{k4}\}$ transmitted on subband k from the second transmit antenna pair.

[1136] For the non-repeated Walsh-STTD scheme shown in FIG. 8, $\hat{s}_{k1}$, $\hat{s}_{k2}$, $\hat{s}_{k3}$, and $\hat{s}_{k4}$ represent the estimates of the four modulation symbols s_{k1}, s_{k2}, s_{k3}, and s_{k4} sent over four transmit antennas on subband k in a 4-symbol period. These symbol estimates may then be multiplexed together into a recovered symbol substream, $\hat{s}_k(n)$, for subband k, which is then provided to multiplexer 1140 in FIG. 11.

[1137] For the repeated Walsh-STTD scheme shown in FIG. 7, one symbol pair $\{s_{k1}$ and $s_{k2}\}$ is sent over both pairs of transmit antennas on subband k in each 4-symbol period. The symbol estimates $\hat{s}_{k1}$ and $\hat{s}_{k3}$ may then be combined by a summer (not shown in FIG. 12) to provide an estimate of the first symbol in the pair, and the

symbol estimates $\hat{s}_{k2}$ and $\hat{s}_{k4}$ may similarly be combined by another summer to provide an estimate of the second symbol in the pair. The symbol estimates from these two summers may then be multiplexed together into a recovered symbol substream, $\hat{s}_k(n)$, for subband k, which is then provided to multiplexer 1140 in FIG. 11.

[1138] For clarity, various details are specifically described for downlink data transmission from an access point to a terminal. The techniques described herein may also be used for the uplink, and this is within the scope of the invention. For example, the processing schemes shown in FIGS. 4, 5, 6, 7, and 8 may be implemented within a multi-antenna terminal for uplink data transmission.

[1139] The MIMO OFDM system described herein may also be designed to implement one or more multiple access schemes such as code division multiple access (CDMA), time division multiple access (TDMA), frequency division multiple access (FDMA), and so on. CDMA may provide certain advantages over other types of system, such as increased system capacity. The MIMO OFDM system may also be designed to implement various processing techniques described in CDMA standards such as IS-95, cdma2000, IS-856, W-CDMA, and others.

[1140] The techniques described herein for transmitting and receiving data using a number of diversity transmission modes may be implemented by various means. For example, these techniques may be implemented in hardware, software, or a combination thereof. For a hardware implementation, the elements (e.g., TX diversity processor, RX diversity processor, TX subband processors, RX antenna processors, RX subband processors, and so on) used to implement any one or a combination of the techniques may be implemented within one or more application specific integrated circuits (ASICs), digital signal processors (DSPs), digital signal processing devices (DSPDs), programmable logic devices (PLDs), field programmable gate arrays (FPGAs), processors, controllers, micro-controllers, microprocessors, other electronic units designed to perform the functions described herein, or a combination thereof.

[1141] For a software implementation, any one or a combination of the techniques described herein may be implemented with modules (e.g., procedures, functions, and so on) that perform the functions described herein. The software codes may be stored in a memory unit (e.g., memory 232 or 272 in FIG. 2) and executed by a processor (e.g., controller 230 or 270). The memory unit may be implemented within the processor or

external to the processor, in which case it can be communicatively coupled to the processor via various means as it known in the art.

[1142] Headings are included herein for reference and to aid in locating certain sections. These heading are not intended to limit the scope of the concepts described therein under, and these concepts may have applicability in other sections throughout the entire specification.

[1143] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

[1144] **WHAT IS CLAIMED IS:**

CLAIMS

1. A method for processing data for transmission in a wireless communication system, comprising:

selecting a particular diversity transmission mode from among a plurality of possible transmission modes to use for each of one or more data streams, wherein each selected diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof;

coding and modulating each data stream based on coding and modulation schemes selected for the data stream to provide modulation symbols; and

processing the modulation symbols for each data stream based on the selected diversity transmission mode to provide transmit symbols for transmission over one or more transmit antennas.

2. The method of claim 1, wherein the plurality of possible transmission modes includes a frequency diversity transmission mode.

3. The method of claim 1, wherein the plurality of possible transmission modes includes a Walsh diversity transmission mode.

4. The method of claim 3, wherein the Walsh diversity transmission mode transmits each modulation symbol over N_T symbol periods, where N_T is the number of transmit antennas used for data transmission.

5. The method of claim 4, wherein the Walsh diversity transmission mode transmits each modulation symbol over all N_T transmit antennas.

6. The method of claim 1, wherein the plurality of possible transmission modes includes a space time transmit diversity (STTD) transmission mode.

7. The method of claim 1, wherein the plurality of possible transmission modes includes a Walsh-STTD transmission mode.

8. The method of claim 1, wherein the plurality of possible transmission modes includes a frequency-STTD transmission mode.

9. The method of claim 7, wherein the Walsh-STTD transmission mode redundantly transmits modulation symbols over a plurality of pairs of transmit antennas.

10. The method of claim 7, wherein the Walsh-STTD transmission mode transmits different modulation symbols over different pairs of transmit antennas.

11. The method of claim 1, wherein the wireless communication system is a multiple-input multiple-output (MIMO) communication system, and wherein the transmit symbols for the one or more data streams are transmitted over a plurality of transmit antennas.

12. The method of claim 11, wherein the MIMO communication system utilizes orthogonal frequency division multiplexing (OFDM).

13. The method of claim 12, further comprising:

OFDM modulating the transmit symbols for the one or more data streams to provide a stream of transmission symbols for each transmit antenna used for data transmission.

14. The method of claim 12, wherein the transmit symbols for each data stream are transmitted on a respective group of one or more subbands.

15. The method of claim 1, wherein at least one data stream is transmitted for an overhead channel.

16. The method of claim 14, wherein the data stream for a broadcast channel is transmitted based on a fixed diversity transmission mode.

17. The method of claim 1, wherein at least one data stream is user-specific and transmitted for a specific receiver device.

18. The method of claim 17, wherein data rate for each of the at least one user-specific data stream is adjusted based on transmission capability of the specific receiver device.

19. The method of claim 1, further comprising:

multiplexing pilot symbols with the modulation symbols for the one or more data streams.

20. The method of claim 1, wherein the pilot symbols are multiplexed with the modulation symbols using frequency division multiplexing (FDM).

21. A method for processing data for transmission in a multiple-input multiple-output (MIMO) communication system that utilizes orthogonal frequency division multiplexing (OFDM), comprising:

selecting a particular diversity transmission mode from among a plurality of possible transmission modes to use for each of one or more data streams, wherein each selected diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof by using frequency diversity, Walsh transmit diversity, space time transmit diversity (STTD), or any combination thereof;

coding and modulating each data stream based on coding and modulation schemes selected for the data stream to provide modulation symbols; and

processing the modulation symbols for each data stream based on the selected diversity transmission mode to provide transmit symbols for transmission over a plurality of transmit antennas.

22. The method of claim 21, wherein the plurality of possible transmission modes includes a frequency diversity transmission mode, a Walsh diversity transmission mode, and a STTD transmission mode.

23. The method of claim 22, wherein the plurality of possible transmission modes further includes a Walsh-STTD transmission mode.

24. A method for processing data for transmission in a multiple-input multiple-output (MIMO) communication system that utilizes orthogonal frequency division multiplexing (OFDM), comprising:

coding and modulating data to provide one or more substreams of modulation symbols for each of a plurality of OFDM subbands; and

for each of the plurality of OFDM subbands, processing the modulation symbols in the one or more substreams for the OFDM subband to provide transmit symbols, wherein the modulation symbols are processed in accordance with a particular diversity processing scheme selected for the OFDM subband to provide diversity in time, frequency, space, or a combination thereof.

25. The method of claim 24, wherein the selected diversity processing scheme for at least one OFDM subband is a space time transmit diversity (STTD) scheme.

26. The method of claim 24, wherein the selected diversity processing scheme for at least one OFDM subband is a Walsh transmit diversity scheme.

27. The method of claim 24, wherein the selected diversity processing scheme for at least one OFDM subband is a Walsh-space time transmit diversity (Walsh-STTD) scheme.

28. A method for processing a data transmission at a receiver of a wireless communication system, comprising:

determining a particular diversity transmission mode used for each of one or more data streams to be recovered from a received data transmission, wherein the diversity transmission mode used for each data stream is selected from among a plurality of possible transmission modes, and wherein each diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof; and

processing received symbols for each data stream based on the diversity transmission mode used for the data stream to provide recovered symbols that are estimates of modulation symbols transmitted from a transmitter for the data stream.

29. The method of claim 28, wherein the plurality of possible transmission modes includes a frequency diversity transmission mode, a Walsh diversity transmission mode, and a space time transmit diversity (STTD) transmission mode.

30. The method of claim 29, wherein the plurality of possible transmission modes further includes a Walsh-STTD transmission mode.

31. The method of claim 29, wherein the plurality of possible transmission modes further includes a frequency-STTD transmission mode.

32. The method of claim 28, further comprising:

demodulating and decoding the recovered symbols for each data stream to provide decoded data.

33. A memory communicatively coupled to a digital signal processing device (DSPD) capable of interpreting digital information to:

select a particular diversity transmission mode from among a plurality of possible transmission modes to use for each of one or more data streams, wherein each selected diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof;

code and modulate each data stream based on coding and modulation schemes selected for the data stream to provide modulation symbols; and

process the modulation symbols for each data stream based on the selected diversity transmission mode to provide transmit symbols for transmission over one or more transmit antennas.

34. A transmitter unit in a wireless communication system, comprising:

a controller operative to select a particular diversity transmission mode from among a plurality of possible transmission modes to use for each of one or more data streams, wherein each selected diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof;

a TX data processor operative to code and modulate each data stream based on coding and modulation schemes selected for the data stream to provide modulation symbols; and

a transmit processor operative to process the modulation symbols for each data stream based on the selected diversity transmission mode to provide transmit symbols for transmission over one or more transmit antennas.

35. The transmitter unit of claim 34, wherein the plurality of possible transmission modes includes a frequency diversity transmission mode, a Walsh diversity transmission mode, and a space time transmit diversity (STTD) transmission mode.

36. The transmitter unit of claim 35, wherein the plurality of possible transmission modes further includes a Walsh-STTD transmission mode.

37. The transmitter unit of claim 35, wherein the plurality of possible transmission modes further includes a frequency-STTD transmission mode.

38. The transmitter unit of claim 34, wherein the wireless communication system is a multiple-input multiple-output (MIMO) communication system that utilizes orthogonal frequency division multiplexing (OFDM).

39. The transmitter unit of claim 38, wherein the transmit processor is further operative to OFDM modulate the transmit symbols for the one or more data streams to provide a stream of transmission symbols for each transmit antenna used for data transmission.

40. An access point comprising the transmitter unit of claim 34.

41. A terminal comprising the transmitter unit of claim 34.

42. An apparatus in a multiple-input multiple-output (MIMO) communication system, comprising:

means for selecting a particular diversity transmission mode from among a plurality of possible transmission modes to use for each of one or more data streams, wherein each selected diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof;

means for coding and modulating each data stream based on coding and modulation schemes selected for the data stream to provide modulation symbols; and

means for processing the modulation symbols for each data stream based on the selected diversity transmission mode to provide transmit symbols for transmission over one or more transmit antennas.

43. The apparatus of claim 42, wherein the plurality of possible transmission modes includes a frequency diversity transmission mode, a Walsh diversity transmission mode, and a space time transmit diversity (STTD) transmission mode.

44. A receiver unit in a wireless communication system, comprising:

a controller operative to determine a particular diversity transmission mode used for each of one or more data streams to be recovered from a received data transmission, wherein the diversity transmission mode used for each data stream is selected from among a plurality of possible transmission modes, and wherein each diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof; and

a receive processor operative to process received symbols for each data stream based on the diversity transmission mode used for the data stream to provide recovered symbols that are estimates of modulation symbols transmitted from a transmitter for the data stream.

45. The receiver unit of claim 44, further comprising:

a receive data processor operative to demodulate and decode the recovered symbols for each data stream to provide decoded data.

46. The receiver unit of claim 44, wherein the plurality of possible transmission modes includes a frequency diversity transmission mode, a Walsh

diversity transmission mode, and a space time transmit diversity (STTD) transmission mode.

47. An access point comprising the receiver unit of claim 44.

48. A terminal comprising the receiver unit of claim 44.

49. A receiver apparatus in a wireless communication system, comprising:

means for determining a particular diversity transmission mode used for each of one or more data streams to be recovered from a received data transmission, wherein the diversity transmission mode used for each data stream is selected from among a plurality of possible transmission modes, and wherein each diversity transmission mode redundantly transmits data over time, frequency, space, or a combination thereof; and

means for processing received symbols for each data stream based on the diversity transmission mode used for the data stream to provide recovered symbols that are estimates of modulation symbols transmitted from a transmitter for the data stream.

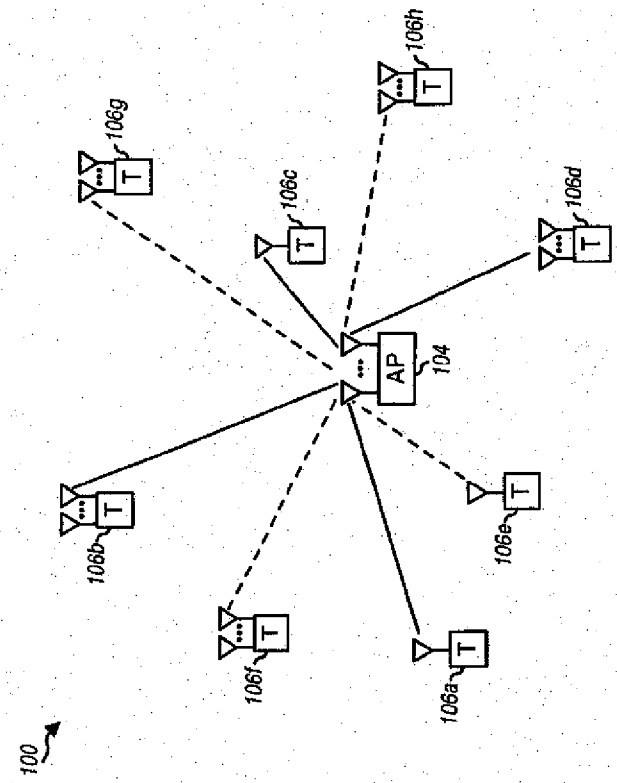
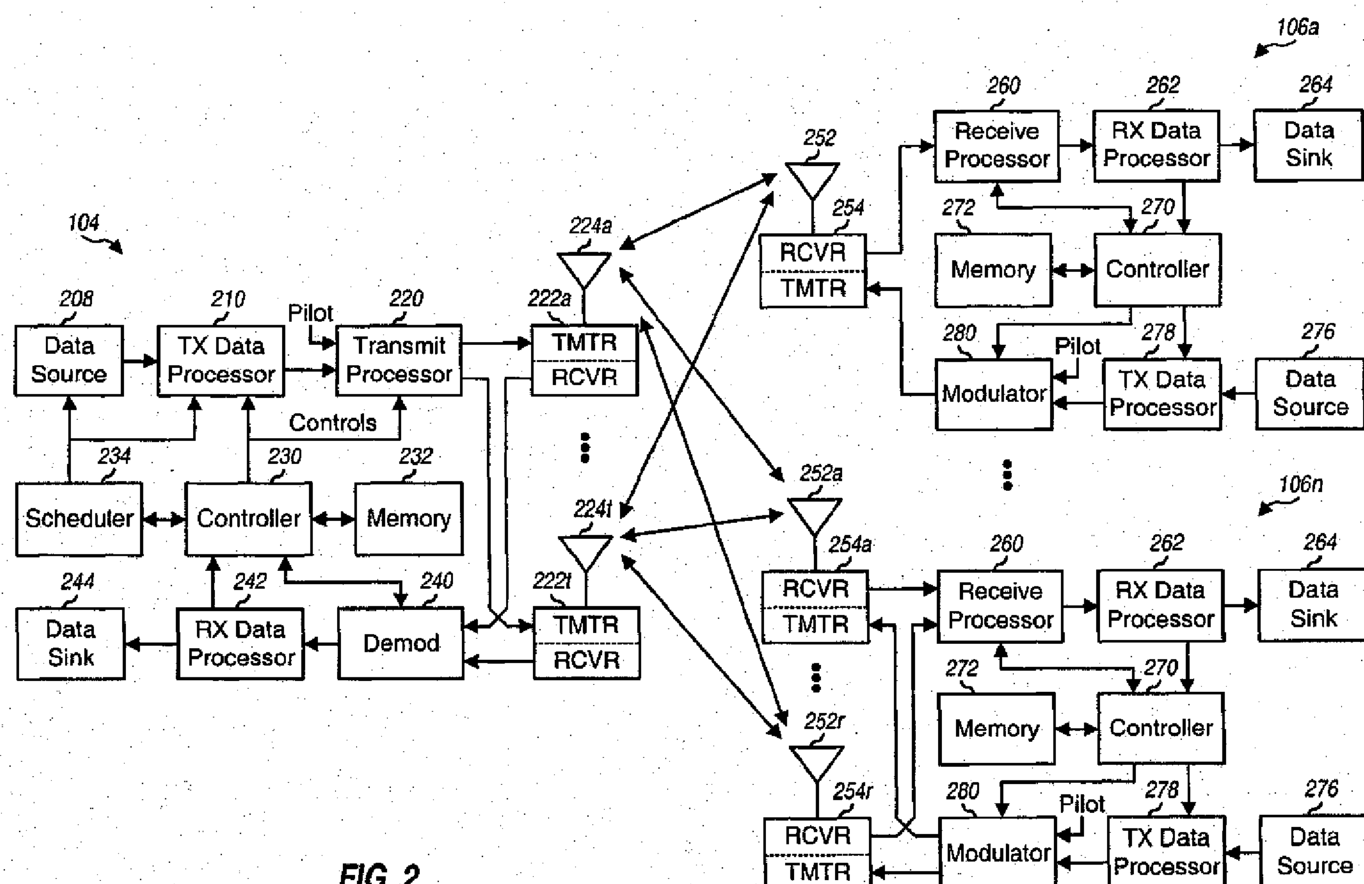


FIG. 1


104
208
Data Source
210
TX Data Processor
Pilot
220
Transmit Processor
222a
TMTR
RCVR
224a
Controls
234
Scheduler
230
Controller
232
Memory
244
Data Sink
242
RX Data Processor
240
Demod
222t
TMTR
RCVR
224t
106a
252
254
RCVR
TMTR
260
Receive Processor
262
RX Data Processor
264
Data Sink
272
Memory
270
Controller
280
Modulator
Pilot
278
TX Data Processor
276
Data Source
106n
252a
254a
RCVR
TMTR
252r
254r
RCVR
TMTR
260
Receive Processor
262
RX Data Processor
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272
Memory
270
Controller
280
Modulator
Pilot
278
TX Data Processor
276
Data Source

FIG. 2

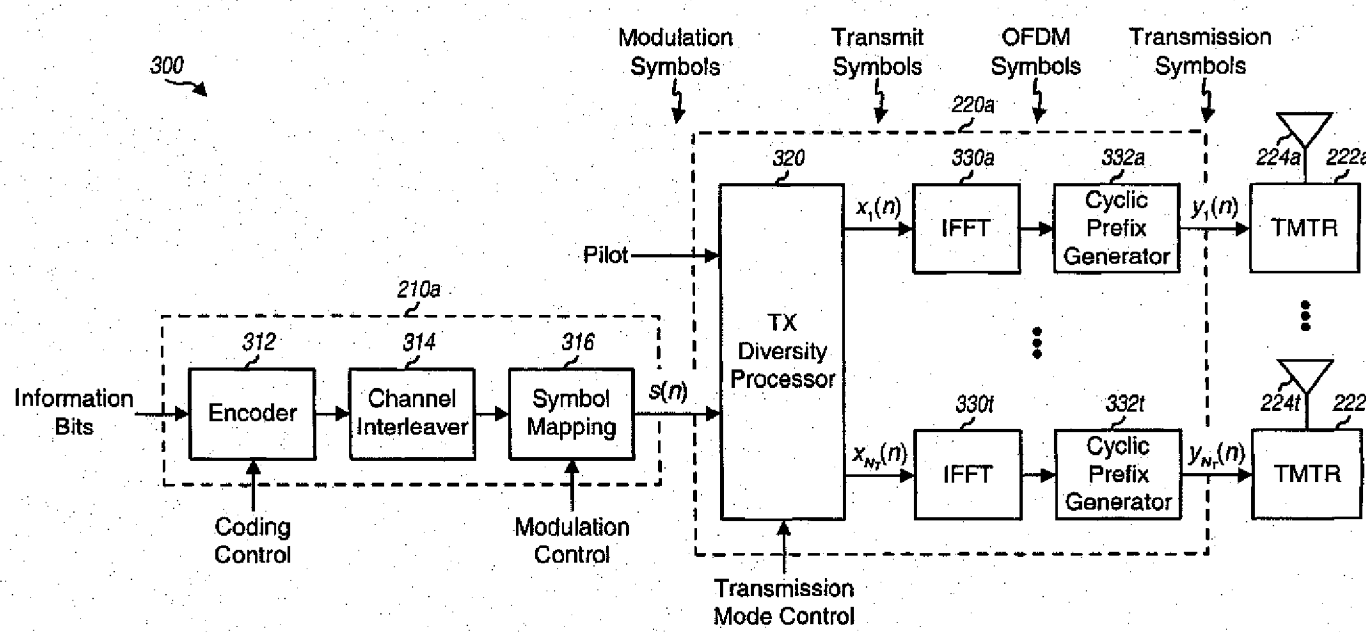
300
Modulation Symbols
Transmit Symbols
OFDM Symbols
Transmission Symbols
220a
320
330a
332a
224a
222a
$x_1(n)$
IFFT
Cyclic Prefix Generator
$y_1(n)$
TMTR
Pilot
TX Diversity Processor
210a
312
314
316
Information Bits
Encoder
Channel Interleaver
Symbol Mapping
$s(n)$
330t
332t
224t
222t
$x_{N_T}(n)$
IFFT
Cyclic Prefix Generator
$y_{N_T}(n)$
TMTR
Coding Control
Modulation Control
Transmission Mode Control

FIG. 3